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Institute of Radio Engineers

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The history of the development of the radio-and-electronic field is packed thick with inspiration and planning, with struggle and achievement. To the workers of today, the past and present of radio offer encouraging pictures and lead to justified hopes for an even brighter and more constructive future.

Presenting in its original form to the readers of the PROCEEDINGS a summary of the historical development of his company's participation in our field, the President of the Stromberg-Carlson Company is lending encouragement to those workers who will likewise be numbered among the leaders in our domain in future years.

*The Editor*

# Radio—Past and Present

WESLEY M. ANGLE

The first time that I personally came in contact with radio engineering was when, around 1907, I read a letter addressed to our company from some yeoman on a naval vessel, stating that certain of the ship's personnel was engaged in experiments in "wireless telephony" and did we have a suitable earcap receiver for use in these experiments. There were several such letters received and we always recommended an operator's headset, with what subsequent good results I have never learned.

Perhaps eight years later, inspired by an advertising manager, who loved to listen to the Arlington time signal, and who tried unsuccessfully one night to arouse in me enthusiasm over the minute sounds I heard, we designed and manufactured a set of headphones, of which we sold an occasional one during the next seven years.

Of course like all members of the reasonably informed general public I had heard of Marconi, had learned of his early feats with wireless, knew that certain ships were equipped with transmitting and receiving apparatus and was stirred by reports of those disasters at sea in which radio began to play a conspicuous part.

But I knew nothing at all of broadcasting until early in 1922, when with our company finally enmeshed in the postwar depression, our purchasing agent began to hear, from his contacts with sundry salesmen, that there was a possible business in radio headsets looming up. In a comparatively short time we had made a few minor changes in our 1915 design and were being flooded with orders. That year we entered orders for 140,000 headsets, manufactured 70,000 and delivered but 35,000. One day in May I prophesied to an associate that the bottom would drop out of the barrel in August. It fell out the following month, that is, in June. A rush of cancellations followed right on the heels of a rush of orders.

Such is the early history of our company's start in the radio business. We brought out our first loudspeaker late in 1923 and our first receiving set in 1924. The part we subsequently came to play in the industry is fairly well known.

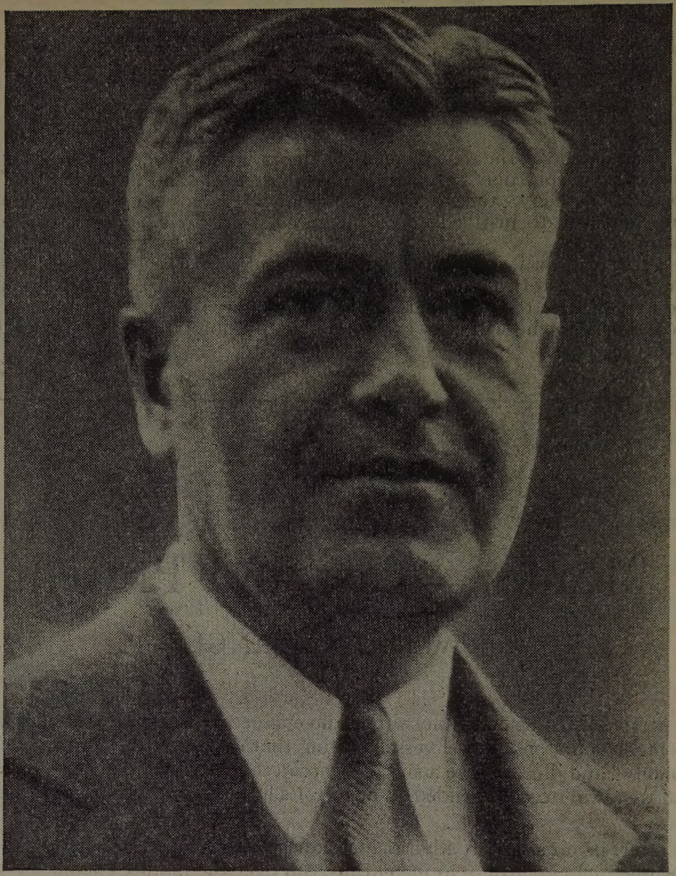
In spite of its very considerable usage on ships at sea, in the transmission of commercial messages and among amateurs, it took about a quarter of a century for radio to emerge from its early beginnings and reach through broadcasting the stage where the lives of millions of people were affected by its usage.

Now another twenty years have passed. Broadcasting has become an accepted part in our routine of living. Ships no longer use radio only to report their positions and send out distress calls but depend on it to determine their positions and thus use it as a means of navigation. Networks for the transmission of messages, largely now having to do with the prosecution of the war, span the globe. And more and more are our lives affected by the use of the many adaptations of Marconi's discovery, not yet fifty years in the past.

And in that comparatively brief span of years, and more particularly in the last ten, the radio engineer has come to be an extremely potent force in our world. I think it fair to state that most radio engineers of twenty years since came to specialize in radio after serving in various other electrical fields. But now the radio engineer has in the main adopted his career right from the time he began his technical training; it is not just his specialty, it is his whole life. And as a consequence there is a constant acceleration of the extent of radio knowledge.

Moreover there is another by no means inconsiderable factor. Radio is being used today in warfare in a tremendous way. And radio technicians are being turned out by the thousands in the schools being maintained by our armed forces. Each one of these men is the potential discoverer of some new application of radio principles.

So I feel that we are today again standing before a new gate which will shortly open to admit us once again into a world where the use of radio for manifold purposes will be far greater than has been the case during the past twenty years. My conclusion is that the radio engineer has chosen a field for his talents second to none in the world.



# Stuart Ballantine

## 1897-1944

Some men, in a relatively brief lifespan, make scientific contributions of outstanding importance. Such a man was Stuart Ballantine, holder of more than thirty patents, author of one book and numerous papers, and recipient of three major awards for distinguished work in the field of radio engineering. A theorist who put his theories into practical service and an inventor whose inventions were of continuing significance in a rapidly developing industry, Stuart Ballantine was a leading personality in his field.

Born on September 22, 1897, at Germantown, Philadelphia, Pennsylvania, Mr. Ballantine's interest in radio developed from a boyhood hobby to a life work. He was educated at Drexel Institute, and in the Graduate School of Harvard University, specializing in mathematical physics. During the first World War he served as an expert radio aide in the Philadelphia Navy Yard, where he had charge of the development of the Navy coil-type compass, a forerunner of radar. He discovered the "antenna effect" in coil-type systems and invented the capacity compensator for its control.

In 1923 he was awarded the John Tyndall Fellowship in physics at Harvard University. At this time he developed the principle of negative feedback to stabilize and reduce distortion in transmission circuits, modulators, amplifiers, and detectors. He was awarded the Liebmann Memorial Prize by The Institute of Radio Engineers in 1931 for his outstanding accomplishments in acoustical and electrical inventions, and in 1934 he received the Elliott Cresson Gold Medal from the Franklin Institute for his work on vertical-antenna radiation.

Mr. Ballantine engaged in extensive studies of detection at high signal levels, fluctuation noise in radio receivers and tubes, development of technique for sound measurements of loudspeakers and receivers, microphone calibration, and broadcast receiver design. He invented a method of stabilizing radio-frequency amplifiers by means of a Wheatstone-bridge circuit, and in 1929 made important contributions to the design and use of vacuum tubes for radio receiving sets, later improving condenser microphones in such a way as to permit increased fidelity in the transmission of sound programs.

One of the most widely known of his many contributions to radio was his invention of the first "throat microphone" to pick up voice sounds directly from the larynx, a device of major importance to aviators, and now widely used by the Army Air Forces.

He founded the Ballantine Laboratories more than ten years ago, and was its president until his death. He was a member of the Franklin Institute, the Radio Club of America, the American Institute of Electrical Engineers, and the American Association for the Advancement of Science. He was also a Fellow of the American Physical Society and the Acoustical Society of America. Mr. Ballantine joined The Institute of Radio Engineers as an Associate in 1916 and transferred to the Fellow grade in 1928. He organized the Philadelphia Section of the Institute in 1920, acting as its chairman until 1926. He served on numerous committees of the Institute, and was its president in 1935.

Mr. Ballantine will long be remembered as a brilliant mathematical physicist, and exceptionally capable engineer, and a gentleman of winning personality.

# The Use of Field-Intensity Measurements for Commercial-Coverage Evaluation*

EDGAR H. FELIX†, SENIOR MEMBER, I.R.E.

**Summary—This paper traces progress in developing methods of defining the commercial-coverage area of broadcast stations by means of field-intensity measurements made by the staff of a service known as Radio Coverage Reports. The initial undertaking of the service was a field survey covering the northeastern states to determine more suitable standards for contour surveys than the accepted 10.0, 2.0, and 0.5 millivolts per meter for urban, suburban, and rural service, respectively. The investigation proved that no fixed set of standard values would serve the purposes of commercial evaluation or would be productive of coverage maps tending to coincide with the disclosures of listener station-utilization findings. The methods used by Radio Coverage Reports in reporting "complete spectrum observations," defining day and night physical delivery of every audible station in cities of over 25,000 population are examined and the resulting concept of a "prevailing standard" based on available service is explained. A correlation of an extensive co-incidental listener survey made by C. E. Hooper, Inc., with Radio Coverage Reports findings shows that the listening audience varies in accordance with the prevailing standard of physical delivery. The possibilities of mapping station and network coverage in accordance with an accurately determined prevailing service are then explored.**

## I. Basis for Evaluating Broadcast-Station Coverage

THE broadcast industry has long needed a really satisfactory method of defining the area served by stations and groups of stations for the requirements of the commercial buyer of radio-advertising facilities. The end result of a satisfactory method must be a simple form of display, preferably a map, suited to statistical tabulation of economic factors, outlining the areas within which a reliable and commercially useful service is offered to the average advertiser.

The wide variety of methods currently used to arrive at a portrayal of coverage depend, in general, upon one of two types of information: (1) technical evaluations secured by field-strength measurements of delivered signal and (2) station- or program-utilization studies based upon tabulation of listener responses to questionnaires, telephone inquiries, or special program offers.

There has been no standardization of terminology or of quantity or quality of research required by either method. The time buyer has been flooded with maps of every type describing primary service, secondary service, listening areas, good service areas, intense service and a score of other designations, none of which has a specific accepted meaning, a minimum or stated standard of field research as its foundation, or submits a result which can be properly compared with the finding of other surveys made for competitive stations. Unified opinion has not developed among time buyers as to the kind of finding desired or the preferred research method on which it should be founded. The varying significance of coverage claims arrived at by different methods is not clearly understood.

In a situation like this, it is natural to look to the engineer for the establishment of a sound technical method of evaluation, free of bias, strictly comparable, and capable of being checked by any competent engineer. The familiar field-intensity survey, showing the 10.0-, 2.0- and 0.5-millivolt-per-meter contours, defining good urban, suburban, and rural service, has not established itself by more than a decade of general use as the satisfactory answer to the requirements of the time buyer.

## II. Attempts to Determine Technical Standards for Coverage Evaluation

After several years of experience in making such field-intensity coverage maps for broadcast stations in all sections of the country, the author became convinced that a more commercially realistic method of coverage definition was needed. Following two years of preliminary research, a service known as Radio Coverage Reports was established in 1935, addressed to finding a solution to this problem. Its first undertaking was an extensive field-research operation in the northeastern states for the purpose of obtaining comprehensive data as to available services in all principal communities of the area with a view to correlating it with the coverage claims issued by stations and networks based on listener research. It was hoped that a more practical set of field-strength standards would be disclosed by analysis of this comprehensive data, the use of which would result in commercial-coverage maps defining areas coinciding closely with the results of experienced advertisers as to the areas served.

The area covered by this pioneer standards survey made in 1933, shown in Fig. 1, was chosen because of the wide diversity of conditions prevailing. It includes every type of terrain from level open farmland of southern New Jersey and Delaware to mountainous sections of New England and New York; numerous concentrated urban areas, including New York, Philadelphia, Baltimore, Washington, and Boston; large and small industrial areas, mining and agricultural sections; in short, the widest possible diversity of conditions and available radio service to be found in an equal area anywhere in the United States. The survey area was bounded by a line drawn north from Washington, D. C., to Syracuse,

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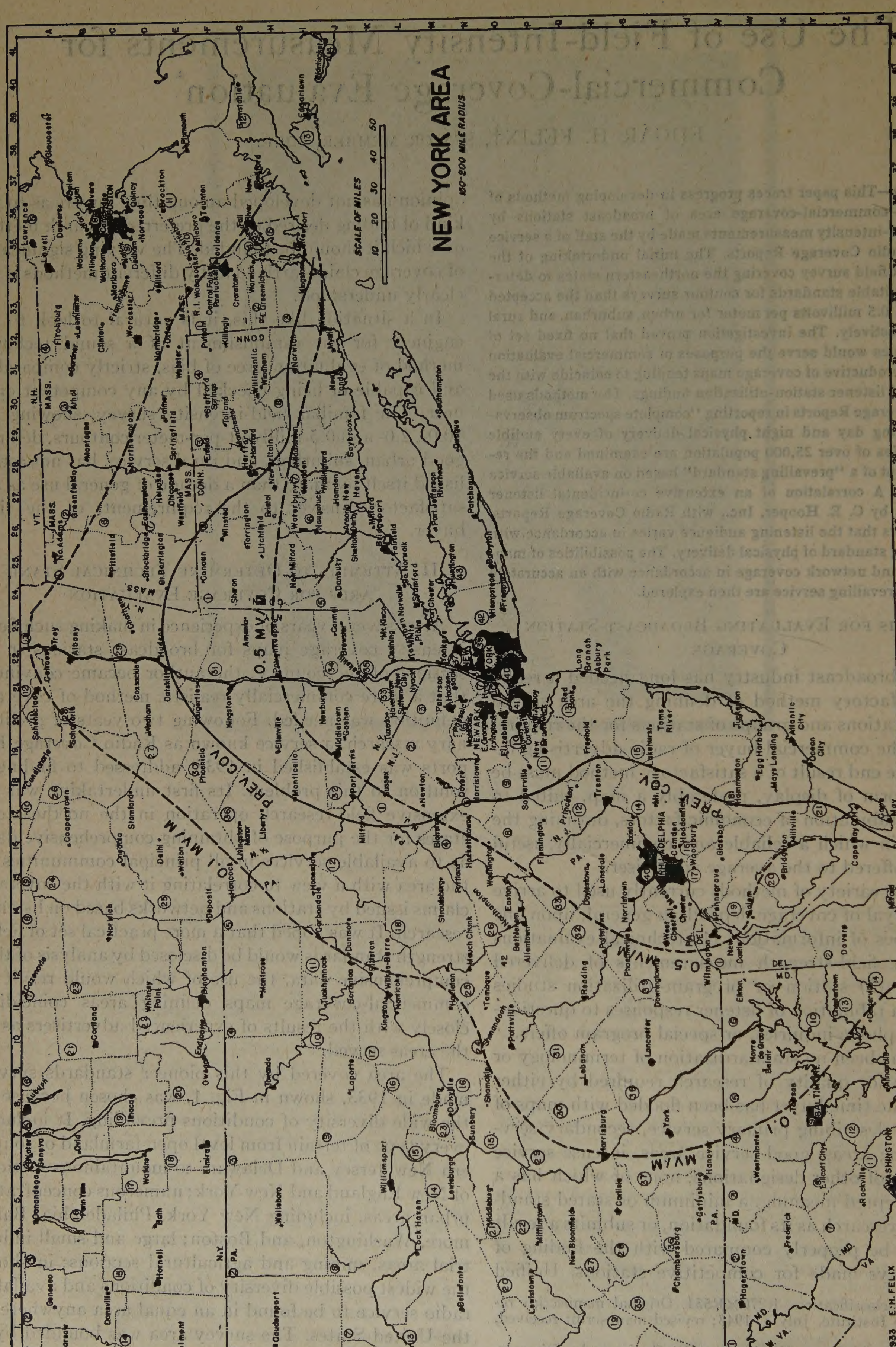



Fig. 1—Prevailing coverage area of a leading New York station, made in 1938, showing 0.5- and 0.1-millivolt-per-meter contours. Note marked departures of coverage from service areas.

TABLE TO ACCOMPANY FIG. 1

CONNECTICUT

| County | No. | Loc. | | Pop. 1930 |
|---|---|---|---|---|
| Fairfield | 5 | K | 25 | 386,702 |
| Hartford | 2 | H | 27 | 421,097 |
| Litchfield | 1 | H | 25 | 82,556 |
| Middlesex | 7 | J | 28 | 51,388 |
| New Haven | 6 | J | 26 | 463,449 |
| New London | 8 | J | 31 | 118,966 |
| Tolland | 3 | G | 29 | 28,659 |
| Windham | 4 | G | 31 | 54,086 |

DELAWARE

| County | No. | Loc. | | Pop. 1930 |
|---|---|---|---|---|
| Kent | 2 | X | 13 | 31,841 |
| New Castle | 1 | W | 13 | 161,032 |

MARYLAND

| County | No. | Loc. | | Pop. 1930 |
|---|---|---|---|---|
| Anne Arundel | 12 | Z | 8 | 55,167 |
| Baltimore | 4 | X | 8 | 124,565 |
| Balto. City | 9 | Y | 8 | 804,874 |
| Caroline | 14 | & | 12 | 17,387 |
| Carroll | 3 | W | 6 | 35,976 |
| Cecil | 6 | W | 11 | 25,827 |
| Frederick | 2 | X | 4 | 54,440 |
| Harford | 5 | W | 9 | 31,603 |
| Howard | 8 | Y | 6 | 16,169 |
| Kent | 10 | Y | 11 | 14,242 |
| Montgomery | 7 | Z | 5 | 49,206 |
| Pr. Georges | 11 | & | 7 | 60,095 |
| Qn. Annes | 13 | Z | 11 | 14,571 |
| Washington | 1 | W | 2 | 65,882 |

MASSACHUSETTS

| County | No. | Loc. | | Pop. 1930 |
|---|---|---|---|---|
| Barnstable | 12 | G | 39 | 32,305 |
| Berkshire | 1 | D | 25 | 120,700 |
| Bristol | 10 | F | 35 | 364,590 |
| Dukes | 13 | I | 38 | 4,953 |
| Essex | 5 | A | 36 | 498,040 |
| Franklin | 2 | B | 27 | 49,612 |
| Hampden | 6 | C | 27 | 335,496 |
| Hampshire | 7 | E | 29 | 72,801 |
| Middlesex | 4 | B | 34 | 934,924 |
| Nantucket | 14 | J | 41 | 3,678 |
| Norfolk | 9 | D | 35 | 299,426 |
| Plymouth | 11 | F | 37 | 162,311 |
| Suffolk | 8 | C | 36 | 879,536 |
| Worcester | 3 | C | 31 | 491,242 |

NEW JERSEY

| County | No. | Loc. | | Pop. 1930 |
|---|---|---|---|---|
| Atlantic | 18 | W | 18 | 124,823 |
| Bergen | 3 | M | 21 | 364,977 |
| Burlington | 14 | T | 18 | 93,541 |
| Camden | 16 | U | 16 | 252,312 |
| Cape May | 21 | Z | 17 | 29,486 |
| Cumberland | 20 | X | 15 | 69,895 |
| Essex | 6 | N | 20 | 833,513 |
| Gloucester | 17 | U | 15 | 70,802 |
| Hudson | 7 | O | 21 | 690,730 |
| Hunterdon | 8 | P | 18 | 34,728 |
| Mercer | 12 | R | 17 | 187,143 |
| Middlesex | 11 | Q | 19 | 212,208 |
| Monmouth | 13 | R | 20 | 147,209 |
| Morris | 5 | N | 18 | 110,445 |
| Ocean | 15 | T | 20 | 33,069 |
| Passaic | 2 | M | 19 | 302,129 |
| Salem | 19 | W | 15 | 36,834 |
| Somerset | 9 | P | 18 | 65,132 |
| Sussex | 1 | L | 18 | 27,830 |
| Union | 10 | Q | 19 | 305,209 |
| Warren | 4 | N | 16 | 49,319 |

NEW YORK

| County | No. | Loc. | | Pop. 1930 |
|---|---|---|---|---|
| Albany | 28 | C | 21 | 211,953 |
| Allegany | 14 | D | 1 | 38,025 |
| Bronx | 37 | N | 22 | 1,265,258 |
| Broome | 22 | E | 11 | 147,022 |
| Cayuga | 5 | A | 8 | 65,751 |
| Chemung | 18 | F | 7 | 74,680 |
| Chenango | 23 | B | 13 | 34,665 |
| Columbia | 29 | E | 29 | 41,617 |
| Cortland | 21 | B | 10 | 31,709 |
| Delaware | 25 | E | 16 | 41,163 |
| Dutchess | 31 | H | 23 | 105,462 |
| Genesee | 1 | A | 1 | 44,468 |
| Greene | 27 | E | 20 | 25,808 |
| Herkimer | 9 | A | 16 | 64,006 |
| Kings | 40 | O | 21 | 2,560,401 |
| Livingston | 2 | B | 1 | 37,560 |
| Madison | 7 | D | 6 | 39,790 |
| Montgomery | 10 | A | 18 | 60,078 |
| Nassau | 42 | N | 23 | 303,053 |
| New York | 38 | N | 21 | 1,867,312 |
| Oneida | 8 | A | 14 | 198,763 |
| Onondaga | 6 | A | 10 | 291,606 |
| Ontario | 3 | A | 4 | 54,276 |
| Orange | 32 | K | 20 | 130,383 |
| Otsego | 24 | B | 16 | 46,710 |
| Putnam | 34 | J | 22 | 13,744 |
| Queens | 39 | O | 22 | 1,079,129 |
| Rensselaer | 13 | B | 23 | 119,781 |
| Richmond | 41 | P | 21 | 158,346 |
| Rockland | 33 | L | 21 | 59,599 |
| Saratoga | 12 | A | 21 | 63,314 |
| Schenectady | 11 | A | 20 | 125,021 |
| Schoharie | 26 | C | 18 | 19,667 |
| Schuyler | 17 | D | 6 | 12,909 |
| Seneca | 4 | B | 7 | 24,983 |
| Steuben | 15 | D | 4 | 82,671 |
| Suffolk | 43 | O | 27 | 161,055 |
| Sullivan | 36 | I | 17 | 35,272 |
| Tioga | 20 | E | 9 | 25,480 |
| Tompkins | 19 | F | 7 | 41,490 |
| Ulster | 30 | G | 20 | 80,155 |
| Westchester | 35 | L | 22 | 520,947 |
| Yates | 16 | B | 5 | 16,848 |

PENNSYLVANIA

| County | No. | Loc. | | Pop. 1930 |
|---|---|---|---|---|
| Adams | 36 | U | 5 | 37,128 |
| Berks | 31 | Q | 11 | 231,717 |
| Bradford | 3 | H | 8 | 49,039 |
| Bucks | 33 | R | 15 | 96,727 |
| Cameron | 6 | K | 1 | 5,307 |
| Carbon | 25 | N | 12 | 63,380 |
| Center | 13 | N | 2 | 126,629 |
| Chester | 39 | U | 12 | 126,629 |
| Clinton | 7 | K | 2 | 32,319 |
| Columbia | 16 | M | 8 | 48,803 |
| Cumberland | 28 | S | 4 | 68,236 |
| Dauphin | 29 | Q | 7 | 165,231 |
| Delaware | 41 | U | 14 | 280,264 |
| Franklin | 35 | U | 2 | 65,010 |
| Fulton | 34 | T | 1 | 9,231 |
| Huntingdon | 19 | P | 1 | 39,021 |
| Juanita | 22 | P | 4 | 14,325 |
| Lackawanna | 11 | K | 13 | 310,397 |
| Lancaster | 38 | T | 10 | 196,882 |
| Lebanon | 30 | R | 9 | 67,103 |
| Lehigh | 42 | P | 13 | 172,893 |
| Luzerne | 17 | L | 11 | 444,409 |
| Lycoming | 8 | K | 6 | 93,421 |
| Mifflin | 20 | Q | 3 | 40,335 |
| Monroe | 18 | M | 14 | 28,286 |
| Montgomery | 32 | S | 14 | 265,804 |
| Montour | 23 | M | 7 | 14,517 |
| Northampton | 26 | O | 14 | 169,304 |
| N'thumberl'd | 15 | N | 7 | 128,504 |
| Perry | 27 | Q | 5 | 21,744 |
| Philadelphia | 40 | T | 15 | 1,950,961 |
| Pike | 12 | K | 16 | 7,485 |
| Potter | 1 | H | 2 | 17,489 |
| Schuylkill | 24 | O | 10 | 235,505 |
| Snyder | 21 | O | 5 | 18,836 |
| Sullivan | 9 | K | 8 | 7,499 |
| Susquehanna | 4 | H | 4 | 33,806 |
| Tioga | 2 | H | 4 | 31,871 |
| Union | 14 | M | 6 | 17,468 |
| Wayne | 5 | J | 15 | 28,420 |
| Wyoming | 10 | J | 11 | 15,517 |
| York | 37 | T | 7 | 167,135 |

N. Y., east from Syracuse to north of the Boston metropolitan area, and southwest from Boston to a point due east of Washington, D. C., along the Atlantic coast.

The following is a brief statistical analysis of the area:

| | |
|---|---|
| Number of cities of 25,000 population or over | 31.1 per cent of total United States |
| Total urban population | 39.9 per cent of total United States |
| Total population | 33.5 per cent of total United States |
| Population density by counties: | |
| Minimum | 28 per square mile |
| Maximum | 85,000 per square mile |

The area may therefore be considered as a cross section of one third of the United States, equally divided between urban and nonurban population. In order to make the available data complete, every audible station was measured in a clear and open residential site of every city of over 25,000 population, which in the aggregate, comprised 48.8 per cent of the total population. of the survey area. In cities of over 100,000 population, several widely separated sites were used for making "complete spectrum observations," which was the term adopted for a field-strength study of every audible station at a given site. Finally, complete spectrum observations were made at intermediate rural points, so that a reasonably accurate approximation could be made for the field strength of any station at any normal site within the entire survey area.

## III. Inconsistency of Technical Field-Strength Standards With Actual Commercial Coverage

The results of this extensive field-survey investigation were tabulated, classified, and compared with station-coverage claims based on all types of surveys issued. The data indicated with complete finality that no set of field-intensity standards defining technically adequate urban, suburban, and rural service was applicable as a commercial-coverage standard for all cities of similar size or to any definable and practical set of conditions. The further the analysis of this varying evidence proceeded, the clearer it became that fixed technical standards of field strength, however carefully chosen, would not serve to define commercially useful coverage. In fact, a single value of field strength frequently fails to define the commercial-coverage boundaries served by an individual station.

A comparison of the coverage area claimed by several leading New York area stations based on comprehensive tabulation of mail over long periods with its field-strength contours showed striking disparities. To the southwest toward Philadelphia, the 2.0-millivolt-per-meter contour was, at that time, a fair approximation of claimed "intense" or "primary" coverage, whether the particular area was rural, suburban, or urban in character; to the north, up the Hudson River Valley toward Albany, the same proportion of listener response or return corresponded to the 0.2 or the 0.25-millivolt-per-meter contour; an even lower standard applied in east central Connecticut; to the northeast in the Hartford area, 1.0 millivolt per meter corresponded to the coverage claims; from this value, it rose to 2.0 millivolts per meter on the shore of Long Island Sound. Similar disparities between technically adequate field strength and the evidence of listener surveys as to actual commercial coverage, prevailed in the Boston area, where such large cities as Lawrence and Lowell extensively utilized

services in the order of 0.5 millivolt per meter or less from Boston stations but listeners then made no substantial use of services from the same stations averaging 1.5 and 2.0 millivolts per meter to the south toward Providence.

One consistency between technical field-strength study and coverage claims based on listener findings became evident as the analysis progressed. High popularity or utilization of coverage in any area was invariably confined to stations having a high ranking field strength. Accordingly, stations were then listed for each survey community in order of field strength, and their standing in various listener surveys were checked off. Regardless of the terminology used in the listener-utilization studies substantially all the "intense," "primary," or "favorite" stations in all the communities were among the first four or five heading the field-strength list. Exceptions were noted but none of the widely used stations were of substantially lower field strength than the first four or five stations listed in order of field strength.

The actual value of field strength of the stations heading the list varied over a wide range according to the prevailing values available at the point in question and bore no relation to size of community or noise conditions. Since the various listener surveys were made on rather divergent and usually undisclosed bases and competitive findings were, therefore, not comparable, no positive rule could be formulated for determining from a list of stations in the order of field-strength ranking at which point delivery values were insufficient to constitute commercial coverage. However, it was evident, because of absence of exceptions, that *listeners rely principally on radio stations offering delivery service of a field strength substantially equal to or exceeding that of the fifth ranking station.*

A further improvement in this correlation between field-strength ranking and useful commercial coverage resulted from classifying stations by groups according to an arbitrary scale of values rated from 1 to 10 rather than in accordance with their precise field strength at a given point in a community. Although hundreds of telephone and mail surveys have since been examined and compared with Radio Coverage Reports ratings in the period from 1936 to 1942, no station in any area has attained its normal proportion of audience (based on audience proportion in high-level areas) where it is of a materially lower Radio Coverage Reports classification than that of the fifth leading station. This applies, regardless of the field strength of the prevailing available service or the number of stations in the same or higher classifications of service.

Fortified by this conclusion, Radio Coverage Reports, previously issued on a regional basis, began to issue a regular service to advertising agencies and broadcast networks early in 1936, reporting the results of field measurements of every audible station for cities of over 25,000 population from coast to coast and for smaller cities having broadcast stations.

## IV. Definitions Applying to Coverage and Service

Because of widespread confusion in terminology, definitions were adopted in connection with the new coverage reporting service, as follows:

### (*a*) *Service Area*

*Service area* is the area within specified contours obtained through field-intensity measurements, representing technically adequate service.

Because of general confusion as to the meaning of "primary" or "secondary" service areas, no attempt was made to attach specific meanings to these associated terms. "Service," in its broadest sense, is a duty or task performed or a readiness to perform it; an adequate service is one which fulfills a requirement. Applied to radio-broadcast-service areas, it indicates only a readiness or capacity to serve within defined areas through the delivery of a field strength adequate to actuate typical radio receivers. Delivery of an adequate service, therefore, carries no implication that it is commercially important or competitively significant.

### (*b*) *Coverage Area*

The *coverage area* is defined as that within which a broadcast station is regularly capable of being utilized by a sufficient proportion of listeners to be commercially significant and valuable as a medium of intelligence and entertainment.

Coverage, therefore, implies the attainment of delivery values competitive with the better available services so that they are normally sampled by listeners seeking entertainment. "To cover," in its general sense, means to place or spread something over, to envelop, to overwhelm; in connection with advertising, *Webster's International Dictionary* defines the term as follows: "to have (a locality or group of persons) as one's territory or field of activity, as in selling merchandise or promoting the interests of a company or rendering a social or business service; as, a salesman *covers* Ohio."

The distinction between service and coverage is important because the terms are too often used loosely. Commercially useful coverage may, in some cases, require several times the field strength necessary to satisfactory service in congested areas; on the other hand, in sparsely served areas, it may be attained by stations delivering far less than technically good service.

A service area can be established only by making field-intensity surveys because it depends upon arbitrary engineering values as its basis. A coverage area can be determined either by field-intensity measurements or by listener studies. An examination of Hooper and Crossley ratings indicates (1) that listeners concentrate most of their attention on a few stations, regardless of the number rendering service and (2) that they fluctuate among these stations from one program period to the next over very wide ratios. If we take as an arbitrary standard, the area over which a station produces

evidence of attaining 5 per cent of radio sets turned on as its audience, that area obtains only during one or more programs. If that station attains an audience of 10 per cent of sets turned on at the limit of its 5 per cent area 15 minutes later, its 5 per cent listening contour is moved further out.

Listening is the result of delivering a satisfactory program, not merely the result of delivery. To obtain the highest audience percentage, not only must delivery not suffer in comparison with that of competitors but presentation of a program of high entertainment value must occur in competition with programs of lower entertainment value among the group of stations which are normally sampled in the area.

Actual coverage at a particular time can be determined only by listener investigations conducted at that particular time. It must be made at all places within the area claimed. This is necessary because a different group of stations often compete in one part of a station's area than in another and the signal-strength ratios among the competing stations common to the whole area very sufficiently to constitute an influential factor in selection. The variables are too numerous to permit a simple definition of coverage and an economic method of establishing the coverage area accurately and comparably by listener investigation.

A field-intensity-measurement method of determining coverage area introduces fewer variables, particularly within the ground-wave areas of stations. By examining all the services at a given point, a group may be selected, in the upper field-strength brackets, which constitutes those delivering their programs with field strengths of the prevailing value and the maximum number of stations which listeners normally sample when seeking entertainment. Further evidence is offered later on, which indicates that a satisfactory and standard way of establishing the prevailing service value can be determined upon. Actual listening percentage, which the time buyer is most desirous of knowing, depends upon the particular program situation obtaining at the moment he places his program.

There is no engineering method of establishing the actual division of the listening audience in an area among its leading physical deliveries. The most that an engineering determination of coverage can do is to lay before the time buyer a map showing the area within which a station's delivery is sufficient to be competitive and commercially valuable if associated with an appropriate program. The engineer can make this finding uniform and comparable for all stations and can do it more effectively than any listener research because he alone can eliminate one of the two factors entering into the determination of the area in which listeners use a station: the highly variable moment-to-moment program situation.

Since Radio Coverage Reports define *all* audible services heard in a community, the prevailing standard of service at the disposal of listeners can be approximated by inspection. The values of field strength defining the ten classifications used for reporting day coverage are based on their effect on ear response and the influence of background noise. These classifications were used for six years by radio time buyers and advertising and network executives and this extended trial suggested no modification of the scale originally adopted, which is given in Table I.

TABLE I

Radio Coverage Reports Day Classifications

| Classification | Field-Strength Limits in Millivolts per Meter |
|---|---|
| 10 | Over 25 |
| 9 | 10–25 |
| 8 | 5–10 |
| 7 | 2–5 |
| 6 | 1–2 |
| 5 | 0.5–1.0 |
| 4 | 0.1–0.5 |
| 3 | 0.020–0.1 |
| 2 | Audible but less than accurately measurable |
| 1 | Reserved for later reclassification |

These ratings proved useful in practice because the arbitrary groupings selected took into account all the important factors entering into useful delivery of program service. The classifications are sufficiently broad to permit representative ratings for most communities. Where service in any city varied over too great a range to fall within a single classification, suitable footnotes furnish additional necessary information. The graphic form of presentation permits indicating whether a service is high or low within its classification. For larger communities, several Radio Coverage Reports are issued by districts of the community, revealing any important coverage deficiency due to unsatisfactory transmitter location, low power, or poor allocation confined to limited sections of the community.

Since there is no listening advantage in substantially all communities in delivering more than 25 millivolts per meter over the entire area, the broad 10 classification served its purpose and, at the same time, eliminates the confusion necessary to reporting the higher field strengths prevailing near transmitters. The spread of field strength in each lower classification becomes quantitatively narrower as the list progresses because signal-to-noise ratios often approach critical values affecting entertainment capabilities in these middle classifications. The lower classifications are again numerically broader because any listeners utilizing such services are necessarily willing to tolerate considerable noise with their program services, eliminating the need for finer discrimination of field-strength values.

For convenience in remembering the classifications, services of over 0.5 millivolt per meter are rated 5 or better and areas within the conventionally measured service contours of 0.5, 2.0, and 10.0 millivolts per meter are represented by minimum ratings of 5, 7, and 9, respectively.

Because of the impossibility of obtaining prolonged night recordings at approximately 500 locations for an average of 30 significant services per location and the rather wide variations observed in night-to-night field

strengths from stations depending on sky-wave delivery, evaluation of night service on a field-strength basis was found impracticable. The ratings used were, instead, based upon the stability and physical quality of signal distribution.

Survey crews were sent progressively over specified areas, with the objective of measuring every daylight field strength in significant locations in all cities of over 25,000 population and at intermediate points sufficient to establish station coverage approximately once each year. After two years of operation, the practice was changed to cover the more congested areas of the country twice a year, measuring at half the locations each six months and the other half approximately six months later. Night field-strength recordings and observations were made of the principal clear-channel stations, intermediate-distance stations, and near-by and local outlets, with a view to classifying night service as shown in Table II.

TABLE II

RADIO COVERAGE REPORTS NIGHT CLASSIFICATIONS

| Classification | Definition |
|---|---|
| A | High-level steady service not subject to fading with a field strength of the day rating |
| B | Satisfactory service of the same order of field strength as the day rating, sufficiently free of crosstalk and co-channel interference to be useful for entertainment, though possibly subject to moderate fading and slight background |
| C | Slow drift fading, no fade-outs of less than 0.5 millivolt per meter lasting more than 5 seconds on nights of good transmission conditions |
| D | Rapid fading, with selective fading and fade-outs frequently observed on nights of average transmission |
| E | Clear service under special conditions, subject to crosstalk or co-channel interference under full night conditions, or having limited service schedule but clearly heard after sunset under particular and limited conditions |
| F | Occasional service under special conditions sufficient to permit identification but subject to crosstalk or interference. |

These rather broad classifications were imposed by economic limitations but they proved practicable because they represent conditions which can be readily understood by time buyers and can be determined from repeated field-intensity recordings and listening and output meter observations. Some border-line cases were found to exist but the large amount of cumulative data obtained over a period of years by careful scheduling of night studies reduced these cases to a negligible quantity. Close track was kept of facilities' changes through Federal Communications Commission notifications and all important new installations were made the subject of special concentrated study. As the resources of the service grew, field investigations were scheduled so as to produce prompt reports of new installations affecting large areas and populations.

## V. Observations on the Application of Radio Coverage Reports

In most of the densely populated urban areas, sufficient A and B service, delivered through so-called ground waves, is at the disposal of the listener to preclude reliance on C and D services. Where less than three or four A and B services are available at night, reliance on lower-grade services, usually in the C and D classifications, is indicated. In order to assist in making use of the information reported, not only were all comprehensive listener-response studies examined but questioning of radio dealers through interviews was carried out by field crews in the more sparsely populated areas. Where entire reliance on C and D service was made necessary by absence of any A and B night service, a considerable knowledge of fading qualities of the important clear-channel stations was generally possessed by dealers and users of sets installed in public places. In such areas, the superior stability of stations in the C classification over the D groups was very generally recognized, a distinction hardly understood in congested areas having sufficient A and B service for their needs. On the other hand, where some, but insufficient, A and B service is available to constitute an adequate choice of night-program sources, listeners tended to use stations in the D classification at such times as they delivered their superior peak levels, rather than the C stations, which rarely reach the field strength of the A and B services. Finally when there were four or more A and B services, there was substantially no familiarity with any C and D services.

Through the seven years that Radio Coverage Reports were issued, the extent of the E area of regional stations tended to fall off substantially, due to the operation of changes in allocation policy. During the earlier years, many eastern coastal regional stations enjoyed good early-evening service for distances up to 100 and, in outstanding cases, to 200 miles, for short periods during the early evening, up to the time that midwestern co-channel stations were transmitting under full night conditions. While the F classification does not generally represent commercially useful service, it is essential to report it because listener response is sometimes submitted as evidence of coverage from areas where the station has a rating as low as 2F (audible but subject to occasional interference).

The transition from relatively extensive regional services to the present limited night areas through growth of high-level co-channel interference, the improvement caused by general use of modern vertical radiators on reduction of rapid-fading areas, the decimation of night-service areas of stations assigned to local channels because of closely adjacent assignments to the same frequency, the increase in adjacent-channel interference by assignment of 50-kilowatt regional powers and nearly equivalent concentration of power through directive operation, and the continued trend toward increased service in already observed areas at the expense of choices in the more sparsely populated rural areas, which took place from 1936 to the end of 1942, is clearly reflected in the record maintained by Radio Coverage Reports.

While Radio Coverage Reports are not directly concerned with the causes of fading, a correlation between barometric conditions and stability of distant services was frequently noted. When the greatest difference in barometric pressure between station and reception point

existed, greater stability in reception from distant clear-channel stations was regularly observed; conversely, when the signal path follows a barometric trough, fading tends to be more severe.

vailing coverage standard is readily determined by inspection. Until listener studies are reported by districts within populous areas and special studies for the purpose are made in smaller cities, precise definition of "prevailing standards" cannot be made. As a general rule of thumb, however, a service is not of outstanding commercial value, regularly and normally capable of retaining a useful proportion of the audience, if five or more stations are of substantially higher classification.

| NIGHT SERVICE | | | | | | STATIONS | | | | | | DAY SERVICE | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| A | B | C | D | E | F | RADIO COVERAGE REPORTS CLASSIFICATION | | | | | | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 10 |
| | | | | | | DAY FIELD STRENGTH AND QUALITY OF NIGHT SERVICE | | | | | | | | | | | | | | |
| HIGH LEVEL STEADY | MODERATE FADING OR INTERFERENCE | SLOW DRIFT FADING | RAPID FADING | CLEAR UNDER SPECIAL CONDITIONS | SUBJECT TO OCCASIONAL INTERFERENCE | CALL LETTERS | KILOCYCLES | NETWORK | NOTE | RCR RATING Day | RCR RATING Night | Audible | .02 to .100 Mv/m | .100 to .500 Mv/m | .500 to 1.0 Mv/m | 1.0 to 2.0 Mv/m | 2.0 to 5.0 Mv/m | 5.0 to 10.0 Mv/m | 10.0 to 25.0 Mv/m | Over 25.0 Mv/m |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WQXR | 1560 | | | 6 | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXX | | | | |
| | | | | —— | —— | WJSV | 1500 | CBS | | 2 | E | —— | | | | | | | | |
| | XX | XXXXX | XXXXX | XXXXX | XXXXX | WHOM | 1480 | | | 6 | B— | XXXXX | XXXXX | XXXXX | XXXXX | XX | | | | |
| | | | | XX | XXXXX | WBNX | 1380 | | | 5 | E | XXXXX | XXXXX | XXXXX | XX | | | | | |
| | | | | XX | XXXXX | WEVD | 1330 | | | 5 | E | XXXXX | XXXXX | XXXXX | XX | | | | | |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WOV | 1280 | | | 8— | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | X | | |
| | XXX | XXXXX | XXXXX | XXXXX | XXXXX | WCAU | 1210 | CBS | | 5 | B | XXXXX | XXXXX | XXXXX | XX | | | | | |
| | | | —— | —— | —— | WHAM | 1180 | Blu | | | D | | | | | | | | | |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WNEW | 1130 | | | 7 | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XX | | | |
| | | —— | —— | —— | —— | WTAM | 1100 | Red | | | C | | | | | | | | | |
| | | | —— | —— | —— | WBAL | 1090 | Red | | 3 | D | —— | —— | | | | | | | |
| | | | —— | —— | —— | WTIC | 1080 | Red | 4 | 3 | D | —— | —— | | | | | | | |
| | XXX | XXXXX | XXXXX | XXXXX | XXXXX | WHN | 1050 | | | 6 | B | XXXXX | XXXXX | XXXXX | XXXXX | XXX | | | | |
| | | | —— | —— | —— | KDKA | 1020 | Red | 4 | | D | | | | | | | | | |
| | | | | | | WINS | 1000 | | 2 | 6 | | XXXXX | XXXXX | XXXXX | XXXXX | XXX | | | | |
| | XXX | XXXXX | XXXXX | XXXXX | XXXXX | WAAT | 970 | | | 7 | B | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XX | | | |
| | | | | | | WPAT | 930 | | | 7 | | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | | | | |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WABC | 880 | CBS | | 8 | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XX | | |
| | | —— | —— | —— | —— | WHAS | 840 | CBS | | | C | | | | | | | | | |
| | | | | | | WNYC | 830 | | | 7 | | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXX | | | |
| | | | —— | —— | —— | WGY | 810 | Red | 4 | 3 | D | —— | —— | | | | | | | |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WJZ | 770 | Blu | | 10 | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XX |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WOR | 710 | Mut | | 10 | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXX |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WEAF | 660 | Red | | 9 | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XX | |
| | | | | | XX | WIP | 610 | | | 5 | F | XXXXX | XXXXX | XXXXX | | | | | | |
| | | | | | XX | WICC | 600 | Blu | | 5 | F | XXXXX | XXXXX | XXXXX | XXX | | | | | |
| XXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | WMCA | 570 | | | 8 | A | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XXXXX | XX | | |
| | | | | | XX | WFIL | 560 | Blu | 4 | 5 | F | XXXXX | XXXXX | XXXXX | XX | | | | | |
| A | B | C | D | E | F | | | | | | | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 10 |

FOOTNOTES: 1. Subject to Flutter during Daylight Hours 2. Has Permit for New Construction 3. Has Permit for Change of Frequency 4. Subject to Adjacent-Channel Crosstalk 5. Minimum Rating Observed in Populous Area 6. Estimated; station not on air during tests

SUPPLEMENTARY SERVICES:

| | | | | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| WCNW | 1600 | 4-F | WDRC | 1360 | 3F | WBRB | 1240 | 4F | KYW | 1060(4)3D | | WPEN | 950 | 4F |
| WWRL | 1600 | 3F | WCAP | 1310 | 3F | WFAS | 1240(1)3F | | WIBG | 990 | 4-F | WEEU | 850 | 3 |
| WBYN | 1430 | 4F | | | | WDEL | 1150 | 3F | | | | WPRO | 630(1)3F | |

RADIO SERVICE: Principally residential community, although 10 microvolts per meter is needed in some sections which are industrial, with a 2- to 5-microvolt-per-meter standard for good reception. Adequate service from the network key stations, WABC, CBS, WJZ, Blue, WOR, Mutual, and WEAF, Red, with WABC trailing slightly here. Among the independents, WOV and WMCA are superior to the many others which can be heard.

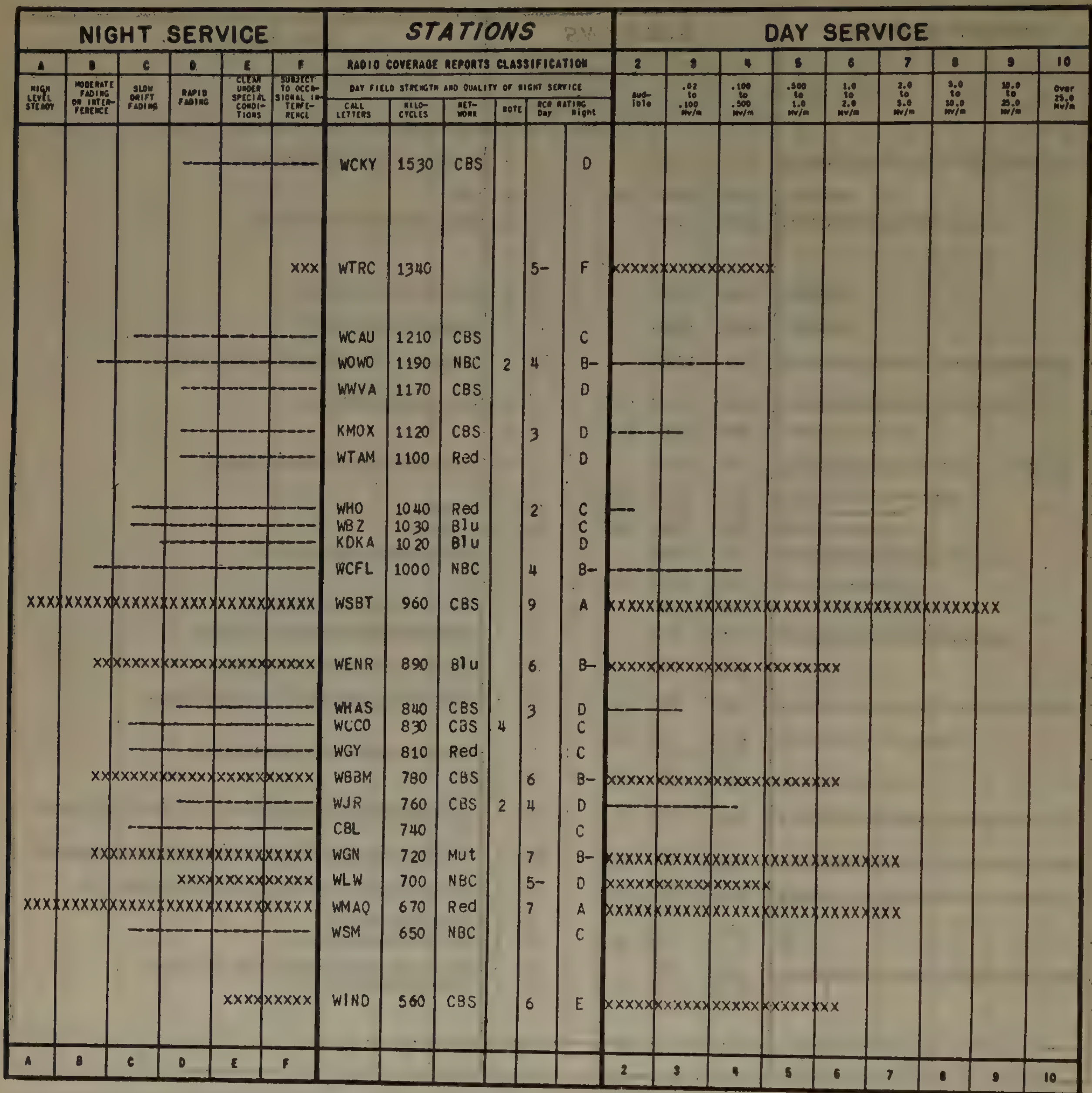
POPULATION AND RADIO FAMILIES: Woodbridge, New Jersey, 25,266 and 5,900
Middlesex County, 212,208 and 49,230 Radio families per thousand population, 232

Fig. 2—A Radio Coverage Report for an overserved area. The prevailing service standard is 8A. Issued in December, 1941.

## VI. Method of Reporting Complete Spectrum Observations

Figs. 2, 3, and 4 show typical Radio Coverage Reports representing widely different types of situations. Because of the graphic method of presentation, the pre-

In an effort to reduce findings to map form, the author prepared many types of special maps based on Radio Coverage Reports ratings, with a view to comparing prevailing service standards with listener surveys. For example, Fig. 1, showing the point at which a leading New York station fell from the fourth ranking to the fifth ranking field strength, made in 1938 (prior to relocation of most of the leading stations of the area), is a reasonable finding of its commercial coverage of the time. However, maps made on this principle for low-power stations have some limitations in overserved areas.




FOOTNOTES: 1. Subject to Flutter during Daylight Hours 2. Has Permit for New Construction 3. Has Permit for Change of Frequency 4. Subject to Adjacent-Channel Crosstalk 5. Minimum Rating Observed in Populous Area 6. Estimated; station not on air during tests

SUPPLEMENTARY SERVICES:

| | | |
|---|---|---|
| WHIP | 1520 | 3 |
| WIRE | 1430 | 3F |
| WIBA | 1310 | 2F |
| WFBM | 1260 | 3F |
| WJJD | 1160 | 4 |
| WRVA | 1140 | E |
| WCBD | 1110 | 4 |
| WIBC | 1080 | 2 |
| WDZ | 1050 | 2 |
| WAAF | 950(4) | 4 |
| WKAR | 870 | 3 |
| CKLW | 800 | 3D |
| WTMJ | 620 | 4E |
| WMT | 600 | 2F |
| WKZO | 590 | 4F |
| WILL | 580 | 4 |
| WKRC | 550(1) | 4-F |

RADIO SERVICE: Typical midwestern industrial city, with 25 microvolts per meter needed near plants, 10 microvolts per meter in most populous residential areas, 2 to 5 microvolt-per-meters satisfactory in outlying districts. Only WSBT, CBS affiliate, offers an adequate service, forcing reliance upon Chicago stations, of which WGN, Mutual, and WMAQ, Red, are the best, with WENR, Blue, WBBM, CBS, and WIND, CBS, trailing but useful.

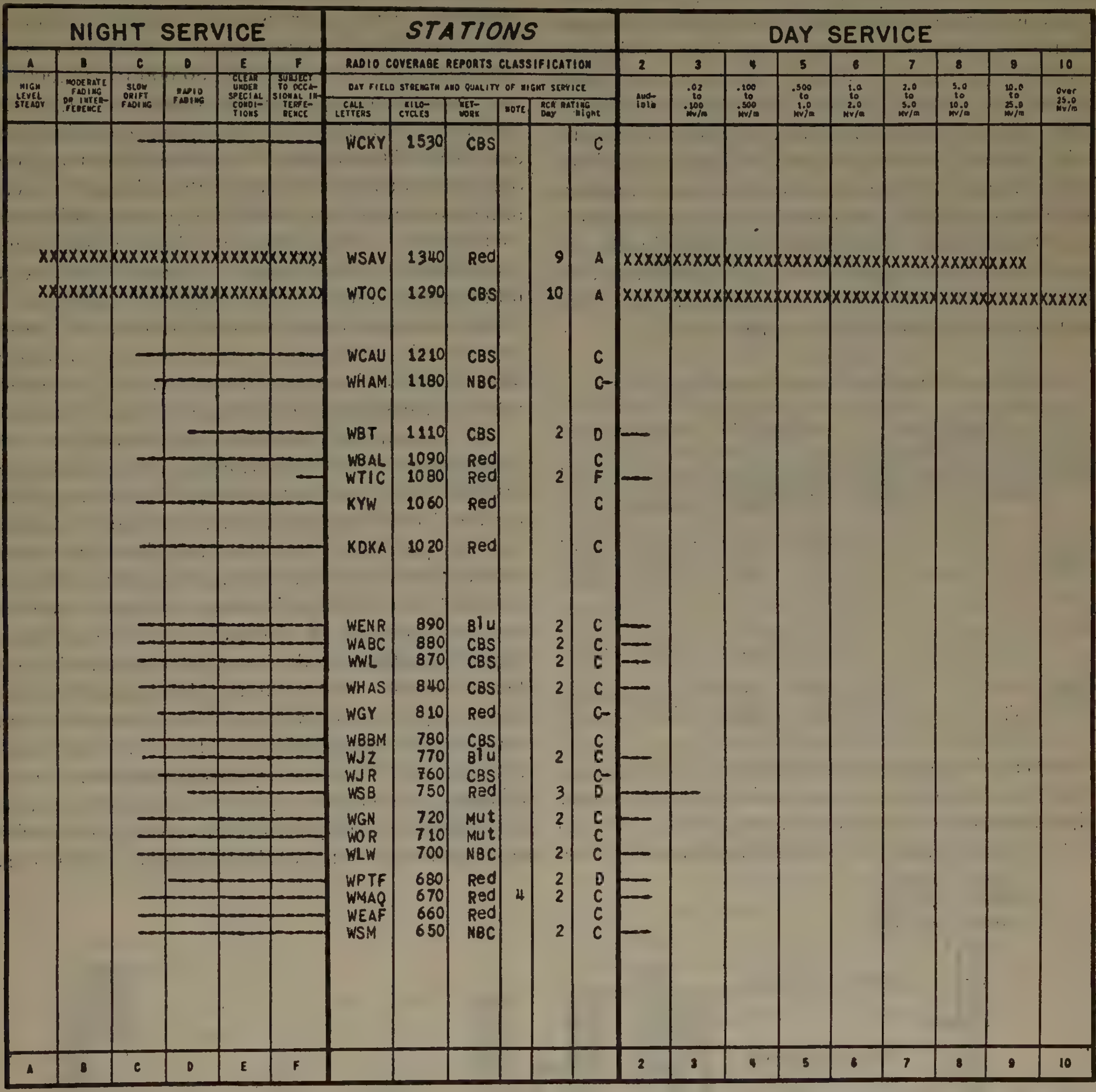
POPULATION AND RADIO FAMILIES: South Bend, Indiana, 104,193 and 25,860. St. Joseph County, 160,033 and 39,710. Radio families per thousand population, 248

Fig. 3—A Radio Coverage Report for a city served from several directions, with a prevailing service standard of 6B. Issued in April, 1941.

The advantage of dial position among the stations having both the highest service levels and the recognized network affiliations is emphasized in the form in which Radio Coverage Reports are prepared. Station ratings are arranged in the order of their frequency and their

field-strength classification and stability ratings are shown graphically, simulating their dial position and relative clarity of reproduction as the listener finds them through the radio receiver in his home.

sections of a metropolitan area and surrounding cities is associated with stations having the highest listener ratings. This follows because facilities attaining superior coverage require substantial investment, excellent cov-




FOOTNOTES: 1. Subject to Flutter during Daylight Hours 2. Has Permit for New Construction 3. Has Permit for Change of Frequency 4. Subject to Adjacent-Channel Crosstalk 5. Minimum Rating Observed in Populous Area 6. Estimated; station not on air during tests

SUPPLEMENTARY SERVICES:

| | | | | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| WALB | 1590 | 2F | WCOS | 1400 | 3-F | WTMA | 1250 | 3F | WMAZ | 940 | 2F | WRUF | 850 | 2F |
| WMOG | 1490 | 4-F | WCSC | 1390 | 4F | WFTM | 1240(1)3F | | WJAX | 930 | 4F | WIOD | 610 | 3F |
| WMFJ | 1450 | 3-F | WJHP | 1310(4)3F | | WAYX | 1230(1)2F | | WGST | 920(1)2F | | WDBO | 580 | 3F |
| | | | | | | WFLA | 970 | 2F | | | | | | |

RADIO SERVICE: The business section and the area surrounding a few large plants calls for 10 microvolts per meter but most of residential section is well served by 1.0 microvolts per meter and can use less under midwinter conditions. In summer 2.0 microvolts per meter is the minimum under most conditions. WSAV, Red, and WTOC, CBS, provide excellent high-level service; all others are inferior by a wide margin. WJAX, Red, WSB, Red, WSM, NBC, and WDVO, CBS, can be heard by day but are very poor. At night, WSAV, Red, and WTOC, CBS, retain their outstanding advantage, although several southern clear channel stations and those in the northeast enjoying over-water transmission come in well at times. Among these are WBAL, WABC, WEAF, WBZ, WSB and WPTF.

POPULATION AND RADIO FAMILIES: Savannah, Georgia, 85,024 and 17,100; Chatham County, 105,431 and 21,150. Radio families per thousand population, 200

Fig. 4—A Radio Coverage Report for a city having insufficient service to provide wide program choice, indicating reliance on C and D service. Issued in November, 1941.

In all large cities having an excess of service, maintained listener surveys are now conducted to permit discrimination of the extent of use which is made of coverage, but even in such places, superior coverage of all

erage attracts excellent programs and the necessary affiliations to produce them; this combination also earns the maximum revenues, permitting investment in the best facilities.

## VII. Correlation Between Relative Field Strength and the Listening Audience Attained

In September, 1938, C. E. Hooper, Inc., specialists in coincidental investigations, co-operated with Radio Coverage Reports by supplying a breakdown of listener distribution among all stations in the New York area. The survey was based on 30,000 telephone calls made within the city limits of New York and in three selected suburban communities, where fairly considerable variations in physical delivery among the stations existed although program service was from the same sources. Radio Coverage Reports are issued for 26 sections within the five boroughs of New York City and in the remainder of its metropolitan area, 32 additional suburban communities are reported, making a total of 58 Radio Coverage Reports for the entire area.

TABLE III

Comparison of Audience Distribution and Signal Strength

| Station and Location | Signal-Strength Ratings | | | Audience Distribution | | Ratio of Signal | |
|---|---|---|---|---|---|---|---|
| | Actual Signal | Prevailing Signal | Technical Standard | Actual Audience | Per Cent of Normal Audience | Per Cent of Prevailing | Per Cent of Technical Standard |
| *Station A* | | | | *Per Cent* | *Per Cent* | *Per Cent* | *Per Cent* |
| Northeast | 6 | 7 | 9 | 27.3 | 89.0 | 86.0 | 67.0 |
| Southwest | 7 | 7 | 9 | 35.7 | 116.0 | 100.0 | 78.0 |
| East | 7 | 8 | 7 | 21.8 | 72.0 | 88.0 | 100.0 |
| *Station B* | | | | | | | |
| Northeast | 6 | 7 | 9 | 10.0 | 62.0 | 86.0 | 67.0 |
| Southwest | 9 | 7 | 9 | 27.5 | 170.0 | 129.0 | 100.0 |
| East | 8 | 8 | 7 | 16.1 | 100.0 | 100.0 | 114.0 |
| *Station C* | | | | | | | |
| Northeast | 9 | 7 | 9 | 48.5 | 145.0 | 129.0 | 100.0 |
| Southwest | 6 | 7 | 9 | 33.2 | 99.0 | 86.0 | 67.0 |
| East | 10 | 8 | 7 | 47.7 | 143.0 | 125.0 | 142.0 |

Table III tabulates the results of this investigation. The field strength of the signals delivered by the three principal stations serving the area are given in terms of Radio Coverage Reports ratings for three near-by communities; the prevailing coverage value is based on the rating of the fifth ranking station at each location and the technical value of field strength required for good service is the established engineering standard for communities of its respective size. The listener-distribution telephone surveys were made simultaneously in the four areas, including two communities to the northeast and southwest, which were in the 50,000 to 100,000 population group requiring at least 10 millivolts per meter for technically good service; the third community to the east of the Metropolitan area, was suburban, requiring only 2.0 millivolts for technically adequate service; the fourth area was New York City proper, which was used to determine average or normal distribution of audience. A reasonably wide range of delivered service values, falling short and exceeding both prevailing and technically required values were to be found in the selected communities, affording an opportunity to observe the effect of such deviations from prevailing and technical values on listening audience percentage attained.

For comparison purposes, the percentage of audience attained at the three locations is compared with normal distribution, based on the audience percentage obtained in New York City itself. Since transmitters are located with the purpose of delivering the highest field strengths at the greatest population concentration within the city limits of New York, this served as a fairly good barometer or standard of normal audience distribution. A better standard would have been obtained if only selected sections within the city limits had been used for listener questioning where all three stations compared served with exceedingly high level of signal, such that listeners would choose their programs solely on the basis of program content, without being influenced by considerations of any physical superiority or deficiency of signal delivery. But, taking the whole of New York City

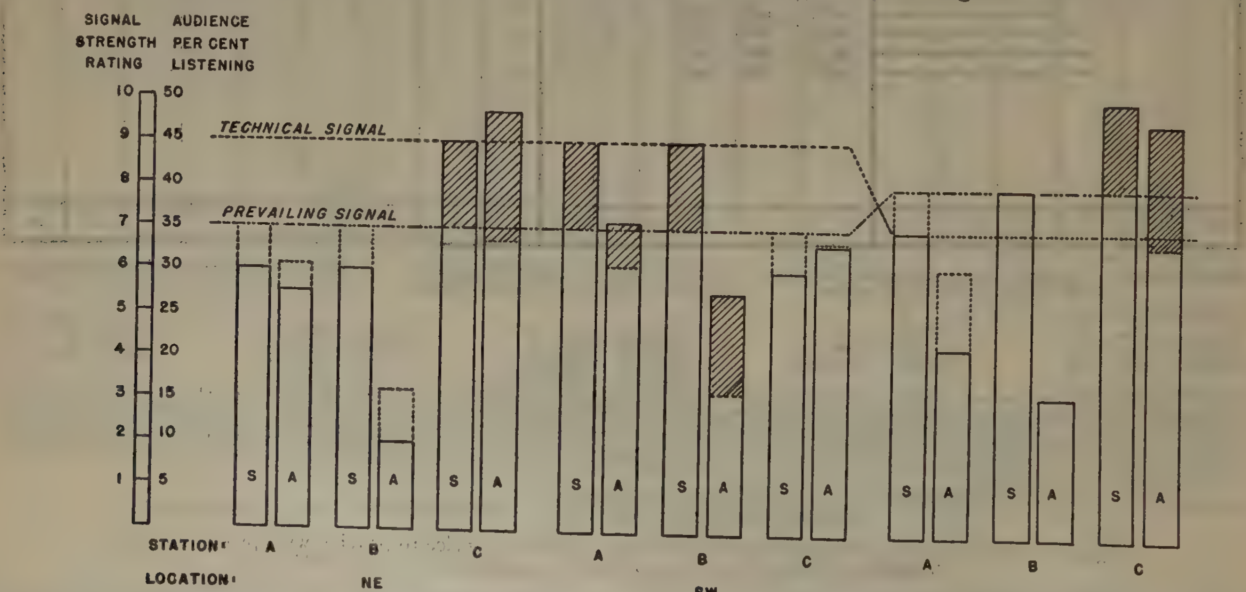



Fig. 5—Relation of field-strength rating to audience attained for three leading New York Area stations in three different sections of the metropolitan area. Solid lines indicate the signal-strength rating (*S*) and the audience percentage (*A*) according to a special Hooper survey. Shaded areas show field-strength ratings above the prevailing value and audience percentages exceeding those attained within New York City limits. Dotted areas show how far below the prevailing coverage value lies and the corresponding loss of audience from the percentage attained in New York City.

tended to cancel out the variations obtaining in some sections, with the result that the base is quite satisfactory. Since all listening determinations were made simultaneously and all four communities depended on the same stations, the only reason for marked differences or deviations from the standard of comparison of audience distribution were due to differences in physical delivery prevailing among the three competing stations in the selected locations.

Table III shows no direct correlation between actual signal and per cent of normal audience attained. For example, Station A attains both its smallest and largest per cent of audience with the same field strength, while Station C has a larger audience return with a weaker signal than in a location with a higher signal delivery. But the correlation between deviations from normal audience and from prevailing signal appears to be quite close.

Fig. 5 shows the percentage of audience attained and the actual rating of the signal delivered in heavy lines for each station and location. In those instances where these quantities fall below the prevailing or normal values, dotted lines are extended to the prevailing signal value and to the normal audience percentage, which is 30.7, 16.1, and 33.3 per cent for Stations A, B, and C, respectively. The shaded areas indicate excess over prevailing signal or normal audience percentage. In every case, failure to deliver prevailing service shows a generally similar proportion of audience loss; exceeding prevailing service always results in attainment of more than the normal share of the available audience.

Fig. 6 emphasizes this correlation. It shows the percentage of normal audience and percentage of prevailing signal delivered for each station and location. A degree of proportionality is indicated, as well as close adherence to the thesis that audience varies in proportion to ratio of delivered signal to prevailing signal.

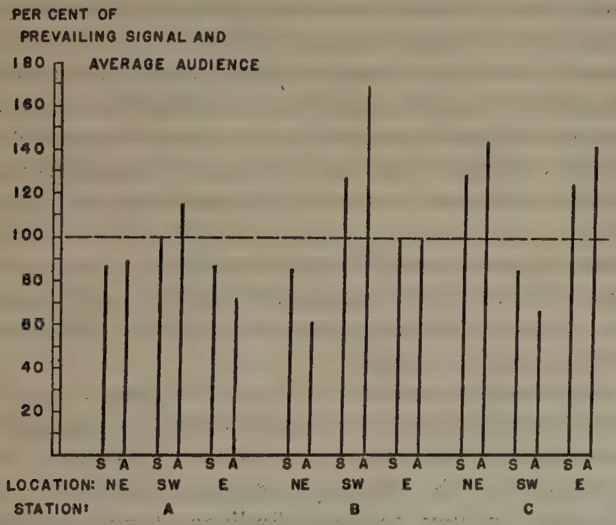



Fig. 6—Percentage of prevailing signal ($S$) and of average audience ($A$), based on prevailing coverage at each point and on Hooper-survey findings for New York City and the community surveyed, arranged in order of location. The audience percentage rises and falls in close consistency with departures from the prevailing service value.

Fig. 7 is a similar comparison with percentage of technically adequate signal, indicating no correlation between technical service and percentage of audience

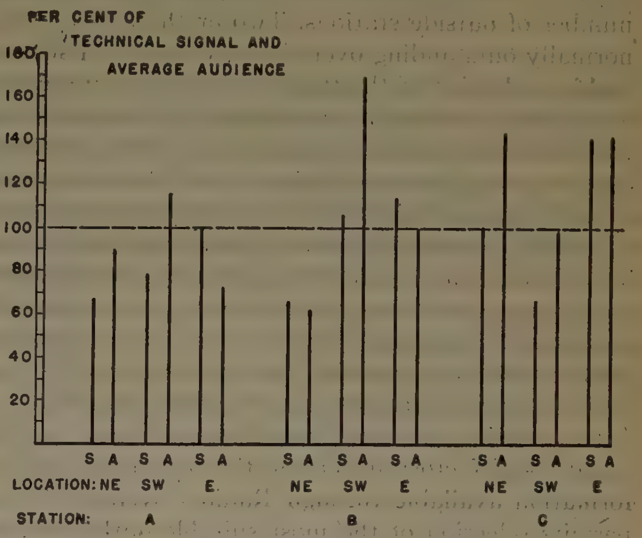



Fig. 7—Percentage of signal delivered to the engineering value necessary to good technical service ($S$) compared with percentage of audience attained as compared with that of New York City proper, according to Hooper. Note the complete lack of any correlation and compare with Fig. 6.

attained. For example, Station A has the smallest percentage of audience in the only location that it attains the technical service standard of delivery; its share of audience is 50 per cent larger at a location where its service is about three fourths of the technical service value. This Hooper survey, therefore, substantiated the findings of the initial Radio Coverage Reports survey that prevailing delivery and not technical engineering standards are the key to commercial coverage area evaluation.

## VIII. Application of Radio Coverage Reports to Commercial Time Buying

Radio Coverage Reports naturally have had their greatest application in rating commercial usefulness of delivery in the hundreds of important communities where sustained listening studies are not conducted. In cities in the 100,000 to 500,000 population group, sufficient locally originated services are generally available to make the principal reliances of the central city fairly obvious but coverage in adjacent communities of such metropolitan areas, the subject of separate Radio Coverage Reports, often varies among the several principal stations to a marked degree. Since the total population of the adjacent cities often exceeds that of the central city of a metropolitan area, Radio Coverage Reports are an important guide in comparing the value of several competing stations in such areas. A station having a high listener rating on the central city where listeners are questioned, may be shown to be incapable of maintaining it in adjacent cities due to delivery deficiencies; or, on the other hand, a station may enjoy even greater popularity in those areas because of delivery deficiencies of its competitors.

In smaller cities, in the 25,000 to 100,000 population group, reliance is generally upon one or sometimes two local or near-by outlets and a choice of a considerable number of outside stations. Two or three of these are normally outstanding over the other nonlocal stations, making selection of the three or four most useful services readily available by inspection. In these situations, there is rarely sufficient standardized information from other sources, particularly since even field-strength surveys are now published without specifying what contour values are measured in arriving at the limits of the area; in any event, surveys for all the competing stations made on the same terms are practically never available for comparison purposes. Radio Coverage Reports have proved useful in evaluating the significance of all types of coverage claims and relating them to each other.

In choosing between two stations which may be equally satisfactory in serving the central city, the information available through Radio Coverage Reports permits selection of the most suitable outlet both for maximum population coverage and for avoidance of duplication of coverage. Sometimes so many cities and stations are involved in making a choice that it becomes necessary to indicate the ratings of all stations under consideration on a map and apply a process of elimination and substitution until a selection of stations affording the maximum coverage within a given budget is determined upon.

The principal limitation in using Radio Coverage Reports is that they are convenient only for city-by-city study of an area and the stations serving it adequately. But they have laid the foundation for further progress in mapping commercially useful areas based on the prevailing service standard.

## IX. Possibilities of Mapping Station and Network Areas to Prevailing Service Standard

Until there are a large number of listener studies available for many communities representing a wide variety of available physical deliveries and types of program services, co-ordinated with such studies in "barometer" areas close to transmitters, no extended study can be made of the direct bearing of deviations from prevailing coverage standards upon actual attained audience. The ultimate objective of utilizing broadcast facilities is the attainment of maximum audience, usually for the purpose of presenting a sales message. It is therefore valid to inquire, since audience surveys are necessary to produce a prevailing service standard, why audience surveys are not a suitable means of determining the entire area served by a broadcast station.

Coincidental listener surveys require that telephone contact be established with an appreciable proportion of potential listeners within an area to determine to what program they are listening each 15 minutes throughout the normal broadcast hours. Only 30 to 40 cities in the United States have enough telephone homes to permit sustained listener surveys to be conducted economically and without repeated annoyance to telephone subscribers. Approximately 200 additional cities are large enough to permit a week's study to be made from one location at reasonably widely separated intervals, from three to six months apart. Even such an additional undertaking would be sufficiently costly to make it doubtful of consummation. It would be relatively easy, given a regular program of investigation, for stations to make unusual efforts, by programming and advertising, to secure an unusually high audience return, during the conduct of a week's special survey in the important cities of its area. Furthermore, it is manifestly impossible to make satisfactory maps upon the resulting findings since they add only a few hundred locations over the whole United States. Consequently, much smaller cities would have to be made the subject of coincidental investigation.

Considerable objection is raised to the validity of coincidental surveys made on the basis of less than 20 completed calls per 15-minute period. Since it takes about four calls to secure one program rating because of radio sets turned off and nonradio homes, 320 calls per hour are needed to make a reasonably accurate coincidental survey. For a 14-hour day and a seven-day week, this means 34,560 telephone calls for a one-week survey. Shorter survey periods are unsuited to this type of finding because the showing of individual stations varies too much from day to day to eliminate the influence of special program situations. Consequently the coincidental type of study is quite limited in its application.

The fixing of a satisfactory standard to constitute a coverage claim on the basis of a week's listening survey involves almost insuperable problems. A minimum standard of 5 per cent average audience attained over all hours would be exceedingly high and would eliminate many stations, with powers as high as 50 kilowatts, in some areas. A minimum standard of 5 per cent for at least one 15-minute period out of each hour would work a hardship on many stations which have morning programs of special appeal and occasionally earn a rating as high as 25 per cent for particular features. Audience percentage variations as high as ten to one from one 15-minute period to the next are not infrequent. Therefore a much lower basis than 5 per cent is impracticable. Any effort to resolve all the variables as to special programs or hours of the day will result in either several definitions of coverage for each station or undue hardship on most of them.

Other survey methods, such as tabulation of mail over long periods, response to special offers or postcard expressions of what stations are used by listeners, overcome some of the objections as to cost and facility, but, in turn, produce coverage findings of more nebulous character which are contributory information but not sufficiently conclusive to serve as an entirely satisfactory coverage definition. Unless a huge and uneconomic quantity of returns are handled, the response per

significant community beyond the immediate central city and its suburbs becomes relatively small. The techniques in this field have been developed through many years of application and the results obtained have contributed materially to general coverage information, but even the most ambitious projects have not produced a specifically defined, comparable, and economic finding, suited to the requirements of positive evaluation. Under the circumstances, the quantitatively firm basis of field-intensity measurements has great advantages as a means of area definition, particularly if the concept of a prevailing service standard can be reduced to practice so that it has a valid relationship to the area in which a station is normally sampled and to comprehensive audience surveys conducted in the principal cities.

Owing to war conditions and the special experience of Radio Coverage Reports personnel useful in military operations, the service has been temporarily suspended but plans for its resumption as conditions permit, in conjunction with coincidental investigations, are well advanced, with a view to establishing a tentative standard for prevailing service in any area and, in turn, to mapping coverage areas for stations and networks on the basis of that standard.

# Equipment and Method for Measurement of Power Factor of Mica*

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***Summary***—**An investigation of domestic sources of mica has been under way for a number of years. The best single electrical property indicative of the suitability of the mica for use in radio condensers is its power factor. Equipment for measurement of power factor, commercially available previous to the present shortage of radio instruments, is described with test procedure.**

## I. Introduction

MICA is one of the strategic materials entering into many phases of the successful prosecution of the war. A very large part of the mica used in this country for radio condensers has been obtained abroad in the past, from India[1] in particular, but this supply has been practically eliminated. It has been desirable therefore to determine whether or not mica from domestic sources and elsewhere in the western hemisphere could be used for condenser purposes in radio equipment. This investigation has been under way for several years by a number of government agencies including the National Defense Council, Bureau of Mines, Geological Survey,[2] War Production Board, and Board of Economic Warfare. The National Bureau of Standards has co-operated in this investigation by making power-factor measurements upon mica from many sources for these agencies.

Since the publication of a paper on mica[3] by the National Bureau of Standards, some notable improvements have been made in equipment for making radio-frequency measurements.

The purpose of this paper is to describe the procedure and equipment now used at the National Bureau of Standards for these measurements, because marked reduction in the time of testing is possible with the new apparatus. This paper is intended to give complete information as to the manufacturers of the equipment used at the Bureau, because of the numerous requests for this information. It will be recognized that some other make of a given device may be used in some cases, but in others there was only one manufacturer of the particular item. The commercial element is therefore unintentional and will be found nonexistent from a practical standpoint, if the purchase of such equipment is attempted. One essential measuring unit is not made now. It is entirely possible, however, that similar testing facilities could be set up if desired, by obtaining these instruments on loan or otherwise, as many have been sold to testing laboratories and broadcast stations.

The power loss in an electrical insulating material is the best single electrical property[4] for indicating the suitability of the material for use as electrical insulation at radio frequencies. The power loss may be expressed in several ways but in many cases is given as power factor in per cent. For some materials the power factor will vary with frequency; in others it may be practically constant with frequency. The lower the power factor the better the material is as an insulator. The best mica has a power factor of 0.02 per cent or less at 100 and at 1000 kilocycles per second.

The measurements on mica[3] reported eleven years ago were made by the resistance-variation method.[4] The

* Decimal classification: R281×R241.5. Original manuscript received by the Institute, August 21, 1943; revised manuscript received, November 16, 1943.

† National Bureau of Standards, Washington, D. C.

[1] Philip R. Coursey, "Electrical Condensers," Sir Isaac Pitman and Sons, Ltd., London, England, 1927, chapter 17, "Mica and mica dielectric condensers."

[2] T. L. Kesler and J. C. Olson, "Muscovite in the Spruce Pine District, North Carolina," Geological Survey Bulletin 936-A, 1942.

[3] A. B. Lewis, E. L. Hall, and Frank R. Caldwell, "Some electrical properties of foreign and domestic micas and the effect of elevated temperatures on micas," Research Paper 347; *Bur. Stand. Jour. Res.*, vol. 7, pp. 403–418; August, 1931. The radio-frequency data in Research Paper 347 are given in summary form. The original data from which the summary was obtained are presented in the Bureau of Mines Information Circular "Mica" by F. W. Horton, I.C. 6822, April, 1935, revised, August 1941.

[4] J. H. Dellinger and J. L. Preston, "Methods of measurement of properties of electrical insulating materials," *Sci. Papers, Bur. Stand.*, no. 471, vol. 19, pp. 39–72; May, 1923.

measurement procedure took about ten minutes when all equipment was functioning satisfactorily. The calculations required at least five minutes. The total time required to determine the power factor of one test specimen at one frequency was therefore about fifteen minutes. With the new equipment the total time required is about three minutes per specimen. The selection and preparation of the specimen for test requires more time than the testing. If splitting of the mica is necessary in the preparation of the test specimen, the total time of preparation may be several times that required for measurement.

## II. Preparation of Specimen for Test

Mica specimens having a thickness between 0.04 and 0.12 millimeter usually have been selected for test. Such thicknesses give capacitances between 800 and 200 micromicrofarads, respectively, with the foil electrodes listed below. An area of the mica free from defects such as pinholes, cracks, bubbles, ridges, broken laminae, spots, blotches, or specks is selected, to which the tin-foil electrodes[5] are attached, thus forming a simple test condenser. Electrodes 19×22 and 25×28 millimeters have been used. The smaller electrode can be approximately centered above the larger one quite easily. The electrodes are fastened to the mica by means of a very small amount of white petroleum jelly, just sufficient to cause the foil electrode to adhere firmly. The electrode is placed on the mica and all air is pressed out by rubbing from the center of the electrode towards the edges with a soft cloth or piece of chamois. Small air bubbles can often be forced out by rubbing with the eraser of a lead pencil, although it is not always possible to eliminate tiny air bubbles completely. If the sample stands a day or two these air bubbles appear to increase in size. Tests have shown that such air bubbles increase the power factor less than 0.005 per cent. The practice is to test the samples soon after preparation. The effect of the tiny amount of adhesive upon the power factor of the mica is nil, as many samples of mica have been tested showing power factors less than 0.01 per cent. If too much adhesive is used, the capacitance of the test condenser seems to be unstable, tending to increase slightly during measurement. This same effect is noted with this equipment when linseed oil is used as the adhesive even in minute quantities. The sensitivity of the test equipment is so much beyond that previously used, that this slight capacitance drift is decidedly detrimental in making rapid measurements.

Fig. 1—Apparatus used in making power-factor measurements upon mica.

## III. Test Equipment

The following test equipment is used in measurements of power factor of mica and is shown in Fig. 1:

Radio-frequency bridge,[6] General Radio Co., type 516-C.

Direct-reading standard condenser,[7] General Radio Co., type 722-N.

Radio-frequency oscillator,[8] General Radio Co., type 484-A.

Radio receiver[9] and speaker, National Co., type HRO.

Special terminals were made and fastened to the "unknown" binding posts of the radio-frequency bridge for holding the mica specimen during measurement. The grounded terminal carried a small coiled spring which placed sufficient pressure on the electrodes of the mica specimen so that no variable resistance was produced at the rounded silver contact points. The mica specimen could be easily placed in position or be removed, in which case the terminals were held apart by means of a small pin and shoulder arrangement.

Experience with the equipment demonstrated that the power-factor adjustment dial and the small condenser dial on the bridge could not be adjusted rapidly as closely as required in these measurements. Accordingly slow-motion devices were attached, but are not shown in the photograph. A belt-and-pulley system was installed to furnish fine control of the power-factor-adjustment dial. A slow-motion device was provided for the small condenser dial. An extension control knob was made for the direct-reading power-factor dial so that the operator's hand would not be near the high-potential connections.

The direct-reading capacitance standard was connected to the "unknown" terminals of the bridge through a link-switching arrangement in the high-potential side of the line. The switch consisted of two mercury-filled cups, with a short copper link to connect the

[5] L. Hartshorn, "Radio-Frequency Measurements by Bridge and Resonance Methods," Chapman and Hall, Ltd., London, England, 1941, p. 192.

[6] C. E. Worthen, "Improvements in radio-frequency bridge methods for measuring antennas and other impedances," *Gen. Rad. Exp.*, vol. 8. pp. 1–6; December, 1933.

[7] D. B. Sinclair, "A high-frequency model of the precision condenser," *Gen. Rad. Exp.*, vol. 13, pp. 1–6; October–November, 1938.

[8] The oscillator shown in the photograph was replaced by a more convenient design described in, "A radio-frequency source for the laboratory," *Gen. Rad. Exp.*, vol. 12, pp. 1–4; November, 1937.

[9] James Millen and Dana Bacon, "Modern design of high-frequency stages for the amateur superheterodyne," *QST*, vol. 19, pp. 13–15; January, 1935.

condenser to the bridge. A special type of small variable condenser was attached to the condenser terminals to aid in rapid precise adjustment of the capacitance. The leads connecting the standard condenser were the same length as those of the special terminals connecting the mica test condenser.

A shielded, constant-frequency oscillator was required to supply voltages at the test frequencies. It was used without modulation as higher precision of measurement can thus be obtained than with modulation.

A communications receiver operating at the test frequencies and capable of receiving unmodulated signals was used with a loudspeaker as the bridge-balance indicator. The local oscillator in the receiving set was adjusted to give an audible frequency of the order of 1000 cycles per second. When the bridge was balanced, the 1000-cycle tone was no longer audible. This form of bridge-balance indicator can often be used successfully in the presence of interference producing another audio frequency, while the bridge-balance indicator mentioned in the next paragraph would not give the desired minimum. A screened room is an essential requirement if much interference is encountered.

A very convenient null balance indicator[5] except at frequencies below 150 kilocycles and between 350 and 550 kilocycles is found in the radio-noise and field-strength meter, model 32A, made by the Ferris Instrument Corporation. This was employed about four years ago and was found to be desirable if the noise from a loudspeaker is objectionable. The instrument includes a radio-frequency voltmeter with a range from 0.1 volt down to the order of a microvolt. A superheterodyne receiver is employed to provide the amplification necessary to operate the output meter.

Concentric leads were used between the radio-frequency bridge and oscillator and detector or receiver. All bridge terminals except the "unknown" or measurement terminals were covered with metal boxes or shields. The table holding the measuring equipment was covered with sheet metal which was grounded. The one "unknown" terminal of the bridge was also grounded. In this manner no difficulties were experienced from capacitance effects from the operator's body.

## IV. Test Procedure

The test procedure consists of three steps: a preliminary capacitance measurement of the mica specimen with the radio-frequency bridge; an adjustment of the bridge power-factor dial to zero with respect to the standard air condenser, and measurement of the power factor of the mica specimen by means of the power-factor dial of the bridge.

## V. Discussion

Many readers are familiar with the American Society for Testing Materials "Tentative Methods of Test for Power Factor and Dielectric Constant of Electrical Insulating Materials, D 150-42 T," and "Tentative Methods of Test for Power Factor and Dielectric Constant of Natural Mica, D 351-42 T," which outline test procedure, method of handling samples, etc. All of the procedures outlined therein have not been followed, particularly that under conditioning of specimens, which provides heating the specimens and keeping them in a desiccator until the specimens are ready for test. Some of the specimens tested have come from the dumps of closed mines, where the samples had been exposed to the weather for many years. Other samples were picked out of pools of water. Many of these samples without any special drying treatment whatever, had as low power factor as the best mica. It therefore seemed unnecessary to follow any special procedure in the treatment of the samples.

The tests at the National Bureau of Standards have not been made to determine changes in power factor with change in pressure of electrodes on the mica. The steel or mercury electrodes specified in ASTM standard D 351-42 T, require two pieces of mica for each test, so that the power factor measured is for two pieces of mica in parallel. A "good" piece of mica measured in parallel with a "poor" piece of mica, will be found to be poor. Hence separate pieces of mica have been used.

The procedure described in this paper is carried out under ordinary laboratory conditions. During the colder months the laboratory temperature is about 23 degrees centigrade and the relative humidity 25 to 30 per cent. During the summer the laboratory temperature may go up to 35 degrees centigrade and the relative humidity may increase to 80 per cent at times. Such changes may affect the measuring instruments and the mica quite appreciably, although the effects are not so significant at the radio frequencies employed as at audio frequencies.

A group of 24 mica samples were tested in August when temperature and humidity were high. The same group of samples was retested in December after the measuring equipment had been exposed to lower humidity which prevails after the heat is turned on in the fall. The mica samples had been stored in paper envelopes in a box. Nineteen of the samples had lower power factors in December than in August. One had the same power factor each time. Twelve of the samples decreased by not more than 0.02 per cent power factor. Seven decreased by from 0.03 to 0.12 per cent power factor. The remaining four samples increased by not over 0.02 per cent power factor.

The power factors of another group of twelve samples were measured as received, after which they were heated for six hours at 149 degrees centigrade. The power factors were remeasured and while they had dropped for ten samples, they increased for two. There was no appreciable change in humidity or temperature of the measuring instruments in this case, but the application of heat to the mica presumably drove off moisture from ten of the samples with an improvement in the power factor. Drying out a dielectric usually lowers the power factor.

[5] *Op. cit.*, p. 99.

The increased values of power factor for the other two samples were found at both 100 and 1000 kilocycles per second. The changes in power factor were of such magnitude as to indicate a change in the mica, and could not be ascribed to errors in the measurements. An explanation of these results might be made if the chemical and capillary structures[10,11] of the mica were known. A limited amount of work has been done seeking to correlate electrical power factor with other physical measurements and observations, but without marked success.

Examination of crystal structure as revealed by X-ray Laue patterns and optical properties as shown by a petrographic microscope failed to offer any explanation for the variation in power factor for the two samples mentioned above or of others having a high power factor.

Tests were made about three years ago upon a number of mica samples to see if there were any relation between the relative clarity of the mica as indicated by its transmission of light and its electrical power factor. No relation was found to exist. Experience gained in testing several hundred samples of mica indicates that color, if uniform, has no definite relation to power factor. One or two cases have been found where the power factor has differed by a factor of three on two areas of a sample an inch or two apart. No visible difference was found. The sample had nothing to indicate a much larger power factor on the one area than on the other. One such sample however showed a double Laue pattern with X rays for the portion having higher power factor. A chemical analysis failed to reveal any marked difference in the two areas.

Mica is sometimes found with many dark or black spots or inclusions of various sizes from a pin point up. Some samples of such appearance have been found to have a low power factor, while other samples have a high power factor. The same may be said for mica with red spots. Many of the black and red stains are probably organic in composition. The mica can often be split and thin laminations be obtained practically free of the colored inclusions.

Another type of inclusion which in the author's experience has always been found to produce a high power factor is rusty brown in color. It often appears in streaks or in larger areas. The power factor of mica with such inclusions may be 1 per cent or higher. These inclusions are probably an iron oxide. Samples having these inclusions have not been split in an effort to find films free of inclusions. A very small amount of such inclusions make the mica unsuited for use in condensers for radio applications.

From the above paragraphs it will be gathered that some mica which is clear and free of inclusions may not be suitable for radio condensers, and that some mica which has inclusions and would not be selected as suitable for radio condensers may have a low power factor. The only way to determine mica suitable for radio-condenser applications so far known is to measure the power factor, rather than judge by appearance alone.

The radio-frequency bridge used in these measurements may be considered as a direct-reading instrument. As most instruments giving direct readings of the quantity measured sacrifice accuracy for convenience in obtaining the answer quickly, it is of interest to know the accuracy of power-factor measurements made with this bridge. A small ceramic-insulated air condenser and a 0.7-ohm straight-wire resistor in series were used as a working standard of known power factor, having a calculated value of 0.11 per cent at 1000 kilocycles per second. Measurements on the bridge gave the same value, while measurements on a twin-T circuit and a $Q$ meter gave 0.12 per cent.

Another working standard of power factor was made up with a silver-mica condenser and 3-ohm resistor. Measurements at 100 kilocycles per second made with other equipment in another Bureau laboratory agreed to 0.01 per cent power factor with measurements on the radio-frequency bridge. This corresponds to half of a division on the power-factor dial, which can be read to 0.002 per cent power factor. Such accuracy is more than adequate for tests upon mica. According to a Bureau of Mines circular[12] good condenser mica should have a power factor not in excess of 0.03 per cent. This limitation could well include a statement of frequency, because some mica has been found to have a much higher power factor at 100 kilocycles per second than at 1000, some samples showing a five-to-one ratio at these frequencies.

Interesting accounts of power-factor measurements elsewhere by means of the $Q$ meter have been published.[13,14]

Some information has appeared recently concerning a previously confidential study arranged by the National Research Council,[15,16,17] of mica and means to determine rapidly its suitability for various electrical uses.

[10] J. W. Williams, "Recent dielectric constant theory and its relation to problems of electrical insulation," *Jour. Franklin Inst.*, vol. 211, pp. 581–606, May, 1931.

[11] R. W. Sillars, "The properties of a dielectric containing semiconducting particles of various shapes," *J.I.E.E.* (London), vol. 80, pp. 378–394; April, 1937; *Proc. Wireless Sect. I.E.E.*, vol. 12, pp. 139–155; June, 1937.

[12] Lawrence G. Houk, "Bureau of Mines Information Circular, 'Marketing Strategic Mica'," I.C. 7219, September, 1942.

[13] H. A. Snow, "Losses in mica and simple test procedure," *Rad. Eng.*, vol. 17, p. 19; February, 1937.

[14] A. W. Barber, "Simplified dielectric loss measurements," *Rad. Eng.*, vol. 17, p. 26; May, 1937.

[15] "Mica and Quartz," (Editorial), *Radio News*, vol. 30, p. 12; November, 1943.

[16] "Spot News, Measuring $Q$ of mica," *FM Radio-Electronics*, vol. 3, p. 60; September, 1943.

[17] F. J. Given, "Mica for war purposes," *Bell Lab. Rec.*, vol. 22, pp. 60–63; October, 1943.

# Copper-Covered Steel Wire at Radio Frequencies*

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*Summary*—At radio frequencies the current in copper-covered steel is confined to the copper portion alone; thus such conductors are electrically equivalent to the corresponding copper tubes. The resistance and internal inductance are as low as for solid-copper wire of the same outer diameter. Without sacrifice of conductance, as much as three fourths of the copper that would be required for a solid-copper conductor can be saved by using copper-covered steel; that is, by the partial substitution of steel. Alternatively, the amount of copper required for a solid conductor can be utilized in copper-covered steel to give a conductor with about half the resistance of the solid-copper wire, and with greater mechanical strength. In general, these observations hold at radio frequencies, but the frequency at and above which they apply depends upon the size of conductor considered. Curves are given to compare the resistance and inductance of copper-covered steel and solid copper over wide ranges of frequency and sizes of conductors.

## Introduction

Copper-covered steel conductors are finding increasing use, chiefly in applications that require strength as well as electrical conductance. In order that the proper conductor may be selected, it is necessary to know the resistance over the frequency range to be employed. Such information has been available only for the very low frequencies and also for the high frequencies where skin-thickness formulas become valid. It is the purpose of this study to determine the resistance of copper-covered steel wire over the whole range of frequencies and to compare it with the resistance of solid-copper wire. The comparisons have especial importance in connection with the saving of copper, a critical war material, by the partial substitution of steel. Such a saving is found to be considerable.

The choice of an electrical conductor involves factors such as conductance, strength, cost, and availability, of which only the first two are considered here. There are in use two types of copper-covered steel conductors with differently proportioned sections of copper and steel having respective nominal conductances 30 and 40 per cent of that of copper wire of the same outer diameter. These figures are minimum specifications; actual conductors generally have conductances exceeding the nominal values by a few per cent. Hence the numerical comparisons to be made hold only approximately for commercial conductors selected at random; discrepancies of the order of a few per cent may be expected. In the 30 per cent conductor the diameter of the steel core is 0.87 of the wire diameter; in the 40 per cent conductor the ratio is 0.80. It follows that the ratios of steel-to-copper area are 3.1 and 1.8, respectively, but since the steel has a resistivity about 15 times that of the copper, the steel core carries only a small fraction of the current and contributes little to the electrical conductance even at low frequencies. For example, at zero frequency, the resistance per unit length of the 30 per cent bimetallic conductor is only 17 per cent lower than that of the copper exterior alone and is only 11 per cent less for the 40 per cent conductor. Thus in the power-frequency range, where copper-covered steel is used for extra strength or for increased diameter to minimize corona losses, the steel core does not alter appreciably the amount of copper required for conductance.

However, at radio frequencies, the situation is different. Here the use of copper-covered steel in most cases permits a saving in copper without loss in conductance. This is brought about by skin effect. The current in the solid copper is distributed unevenly, tending to flow along the surface, and the interior copper is used ineffectively or not at all. The copper-covered steel, on the other hand, behaves like a tube and, at high frequencies, is as good electrically as a solid conductor of the same material and same outside diameter, and, indeed, at some frequencies is even slightly better. Thus, at radio frequencies, copper may be used more effectively in copper-covered steel than in copper to provide electrical conductance. In what follows, the comparisons are made on a quantitative basis.

## Mathematical Expression for Resistance

The alternating-current resistance of copper-covered steel wire is calculated by means of the following equation which was derived elsewhere:[1]

$$\frac{Z}{R_0} = \frac{br_2}{2j} P \operatorname{conj.}\left[\frac{A-(F/H)C-IGA+(IE/H)C}{B-(F/H)D-IGB+(IE/H)D}\right]$$

in which the nine quantities $A$ to $I$ are either Bessel functions expressed as complex numbers, or are quantities simply obtained from Bessel functions. The forms employed were selected for convenience in using the Jahnke-Emde tables.[2] The symbols are defined as follows:

| | |
|---|---|
| $r_1$ = interface radius | $r_2$ = outer radius |
| $\rho_1$ = resistivity of core | $\rho_2$ = resistivity of exterior |
| $\mu_1$ = permeability of core | $\mu_2$ = permeability of exterior |
| $\omega = 2\pi f$ | $j = \sqrt{-1}$ |
| $a = \sqrt{\omega\mu_1/\rho_1}$ | $b = \sqrt{\omega\mu_2/\rho_2}$ |
| $A = J_0(j^{1/2}br_2)$ | $E = J_0(j^{1/2}br_1)$ |
| $B = j^{1/2}J_1(j^{1/2}br_2)$ | $F = J_1(j^{1/2}br_1)$ |
| $C = H_0^{(1)}(j^{1/2}br_2)$ | $G = \dfrac{H_0^{(1)}(j^{1/2}br_1)}{H_1^{(1)}(j^{1/2}br_1)}$ |

[1] B. R. Teare, Jr., and Josephine R. Webb, "Skin effect in bimetallic conductors," *Trans. A.I.E.E.* (*Elec. Eng.*, *June, 1943*), vol. 62, pp. 297–302; June, 1943.

[2] E. Jahnke, F. Emde, "Tables of Functions," B. G. Teubner, Leipzig, Germany, 1933 edition.

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$$D = j^{1/2}H_1^{(1)}(j^{1/2}br_2) \qquad H = H_1^{(1)}(j^{1/2}br_1)$$

$$I = \sqrt{\frac{\mu_2\rho_2}{\mu_1\rho_1}}\frac{J_1(j^{1/2}ar_1)}{J_0(j^{1/2}ar_1)} \qquad P = 1 - [r_1/r_2]^2[1-(\rho_2/\rho_1)].$$

The meter-kilogram-second system of units is employed.

The real component of $Z/R_0$ is the ratio of alternating- to direct-current resistance. The imaginary component is the ratio of the internal reactance to direct-current resistance, the internal reactance being that part of the reactance arising from the flux within the conductor.

The equation applies to any bimetallic conductor of circular section, including aluminum-covered steel as well as copper-covered steel. The quantity $P$ in the equation is the ratio of the conductance of the given conductor to that of a solid homogeneous conductor of the same diameter made of the outer material whatever it may be. Thus for copper-covered steel, $P$ is the per cent conductance divided by 100. By dropping certain terms in the equation and making appropriate changes in $P$, the equation also gives the ratio $Z/R_0$ for solid and hollow homogeneous conductors. Thus, if only the first terms $A$ and $B$ within the brackets are used, and if $P$ is set equal to unity, the expression applies to a solid homogeneous wire of outer radius $r_2$, resistivity $\rho_2$, and permeability $\mu_2$; that is, with the characteristics of the outer material. Likewise, if only the first two terms of the numerator and the first two terms of the denominator are used with a value of $P$ corresponding to an infinite value of $\rho_1$, the expression applies to a tubular conductor of inner and outer radii $r_1$ and $r_2$ made of the outer material.

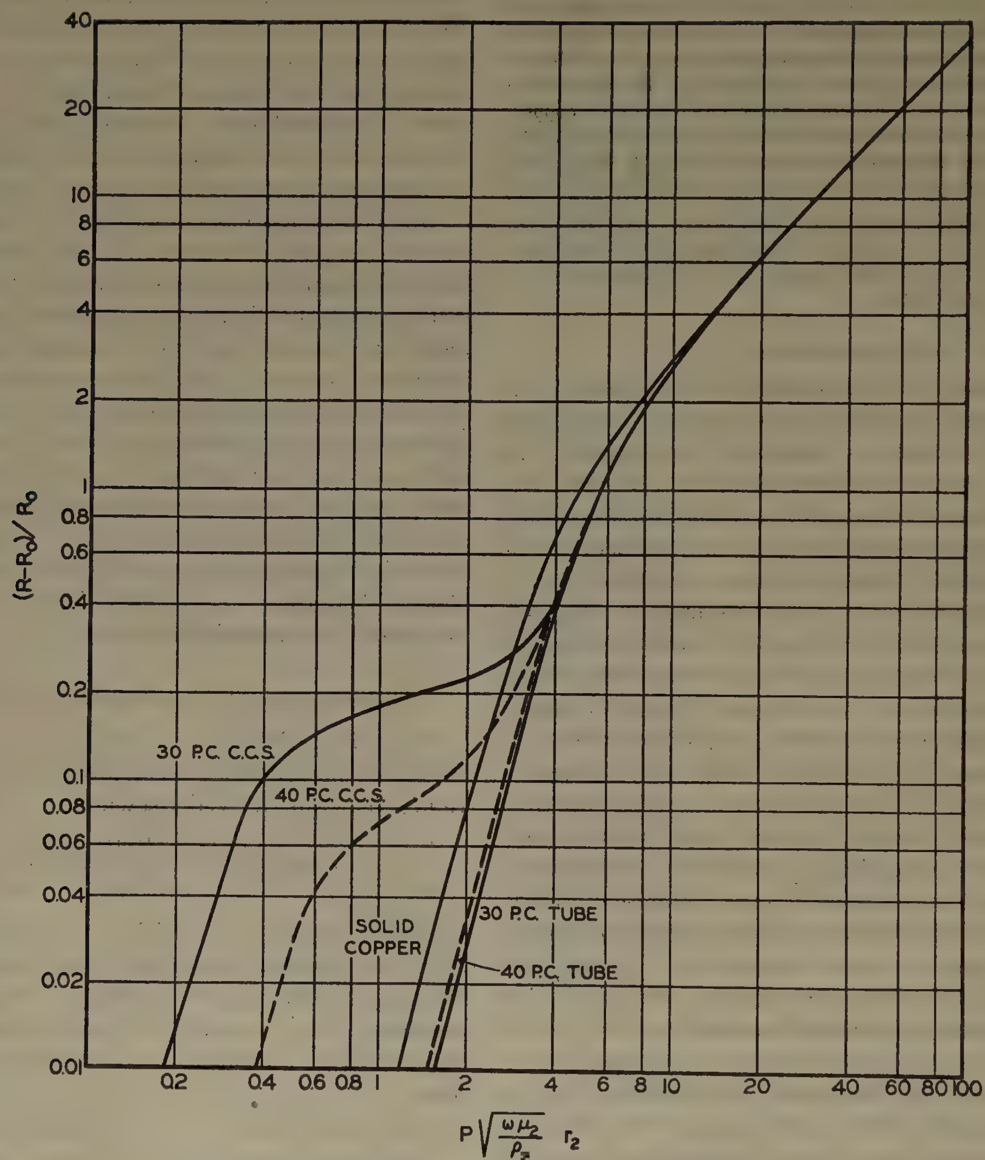


Fig. 1.—Per unit increase of resistance as a function of frequency and outer radius.

The equation was derived on the basis of the following assumptions:

1. The steel and copper regions are homogeneous; that is, within each the permeability and resistivity do not vary from point to point.

2. At any point the permeability and resistivity do not vary with time.

3. The bounding surface between regions as well as the outer surface is a circular cylinder.

4. Current in the conductor varies sinusoidally with time.

5. The conductor is very long and straight, and the return conductor or conductors are either concentric tubes or are at such remote distances that they produce no appreciable field within the conductor considered.

The equation was checked experimentally in the frequency range of most interest, the range in which there is appreciable current in all parts of the copper. Above this frequency range the current approaches a distribution which is confined to a thin skin at the surface, and the equation gives values in agreement with those obtained from the generally accepted expressions for solid conductors.

The curves of Fig. 1 give in a general manner the calculated increase of resistance with frequency for copper-covered steel conductors of nominal 30 and 40 per cent conductance. The actual conductances were taken as 27.7 and 42.3 per cent, respectively, the latter being chosen to conform to a specimen that was tested and the former to simplify the use of the tables of Bessel functions in making the calculations. For purposes of comparison, curves are also included for copper tubes and solid copper. The tubes correspond to the copper exteriors of the copper-covered steel conductors. The ordinates of each curve give per unit increase of resistance over that at zero frequency; the abscissas are in terms of $Pbr_2$, which for copper at 20 degrees centigrade with resistivity $1.77\times10^{-8}$ ohm-meter (10.6 ohms per circular mil-foot) becomes $21.1Pr_2\sqrt{f}$. These abscissas are chosen for generality; that is, to make the curves apply to conductors of any diameter, and also to bring the curves for all conductors together in the region of highest frequencies. In terms of $Pbr_2$ the curves for solid and tubular conductors hold for other materials as well as copper, providing the tubes have similar ratios of inner to outer diameter. Plotting per unit increase of resistance as a function of $Pbr_2$, with both scales logarithmic, yields curves which asymptotically approach straight lines at both high and low extremes of frequency. At the lower end the asymptote has a slope of 4; at the upper end, 1.

## Comparison of Conductors

Copper-covered steel and copper conductors are compared over a large frequency range in three ways; namely, in terms of resistance when the outside diameters are equal, in terms of resistance when both conductors have equal weights of copper per unit length, and in terms of diameter when they have equal resistance. Fig. 2 gives the resistances of the five conductors of Fig. 1 when all have the same outer diameter. The ordinates for any curve give the ratio of alternating-current resistance of the corresponding conductor to direct-current resistance of the solid copper. The abscissas are proportional to the square root of frequency, two scales being given, one directly in frequency for No. 18 conductors (0.040 inch in diameter), and the other generalized for any size of conductor. At the lowest frequencies, near zero, the resistances are necessarily inversely proportional to the per cent conductances. As frequency is increased, the resistance of the solid-copper conductor increases faster than that of

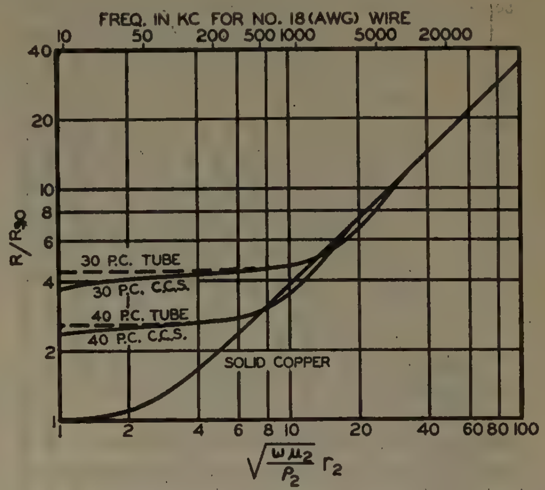


Fig. 2—Ratio of resistance to zero-frequency resistance of solid copper. Conductors all have same diameter. Upper scale shows frequency for No. 18 wire, lower scale for any size.

the copper-covered steel and overtakes the resistance of each copper-covered-steel conductor. The intersection marks the beginning of a frequency range in which the copper-covered steel actually has a slightly lower resistance than solid copper of the same diameter. Finally with a further increase of frequency, the curves for copper-covered steel again approach the curve for solid copper, and all the resistances, including those for the tubes, become the same. The curves also show that except at low frequencies the copper-covered steel and the corresponding tubular copper exteriors have substantially the same resistance.

For many practical problems this last observation suggests a simple means of determining the performance of copper-covered-steel conductors, since data for tubular conductors are available.[3,4]

Fig. 3 gives a ready means of applying the general scale of abscissas $br_2$ of Fig. 2 and the following figures to a copper or copper covered steel conductor of any size. Frequency is plotted as a function of $br_2$ for a large number of conductor sizes from the equation $br_2=21.1r_2\sqrt{f}$ which applies to a copper of resistivity $1.77\times10^{-8}$ ohm-meter. For example, suppose it is desired to find the resistance per 1000 feet of No. 22, 40 per cent conductance copper-covered-steel wire at 0.8 megacycle. Fig. 3 should be used to determine the value of $br_2$. The value found is approximately 6. In Fig. 2 the ratio $R/R_{s0}$ for $br_2$ equals 6 is seen to be approximately 2.8. Multiplying this ratio by the resistance

[3] A. W. Ewan, "A set of curves for skin effect in isolated tubular conductors," *Gen. Elec. Rev.*, vol. 33, pp. 249–252; April, 1930.

[4] L. F. Woodruff, "Electric Power transmission and Distribution," John Wiley and Sons, Inc. New York, New York, 1938.

per 1000 feet of solid copper (16.1 ohms) gives the desired result, 45.1 ohms. In order to use Fig. 3 with Fig. 1 the value of $br_2$ from Fig. 3 must be multiplied by $P$, the per unit conductance, to obtain the abscissas of Fig. 1.

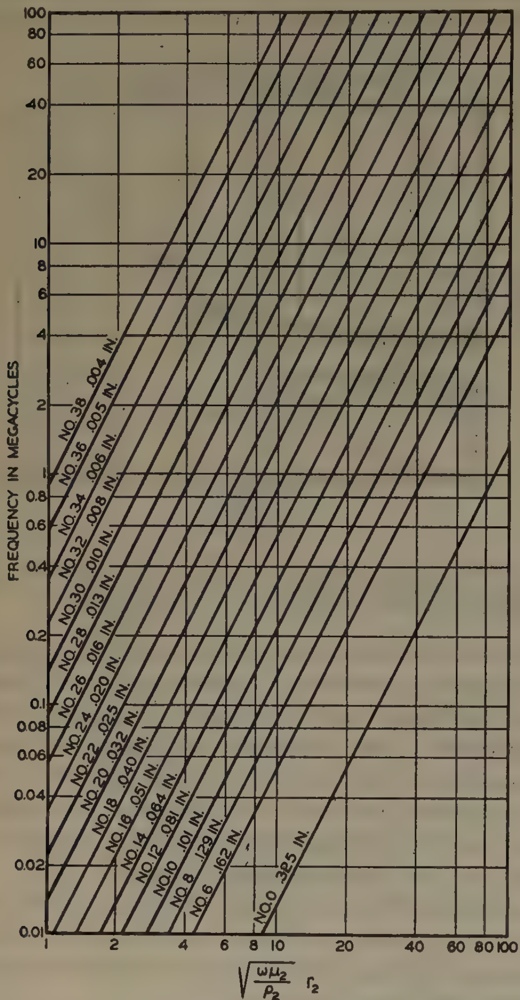


Fig. 3—Chart relating parameter of generalized abscissas $r^2\sqrt{\omega\mu_2/\rho_2}$ to frequency and wire size.

Fig. 4 gives a direct comparison of the resistances of the copper-covered steel conductors with that of a solid conductor of the same diameter and at the same frequency. No. 18 copper-covered-steel wire of 30 per cent conductance reaches equality with No. 18 solid copper at 1710 kilocycles; No. 18 wire of 40 per cent conductance reaches equality at 530 kilocycles, as is shown in the figure. The corresponding equality points for No. 10 wire would occur at 268 kilocycles for 30 per cent conductance and 83 kilocycles for 40 per cent conductance or in general when

$$f = 0.448/r_2^2 \text{ for 30 per cent wire}$$
$$f = 0.139/r_2^2 \text{ for 40 per cent wire}$$

$f$ being expressed in cycles, $r_2$ in meters.

Since these comparisons of copper-covered steel and solid-copper wire are on the basis of equal outside diameters, it is evident that at these frequencies copper could be saved by using it in copper-covered steel rather than in solid conductors. For example, suppose that at a frequency beyond the equality point a given solid-copper conductor has a suitable resistance and mechanical strength. If this solid conductor were replaced by copper-covered steel of the same diameter, the resistance is the same, the strength is considerably increased, and copper of an amount corresponding to the core size is saved, being replaced by the steel. The saving amounts to about three fourths of the copper in the solid conductor for replacement by a 30 per cent conductor and about five eighths for replacement by a 40 per cent conductor.

In Figs. 2 and 4 the basis for the comparison of the copper-covered steel with the solid conductor has been the equality of outside diameters. Consider now the

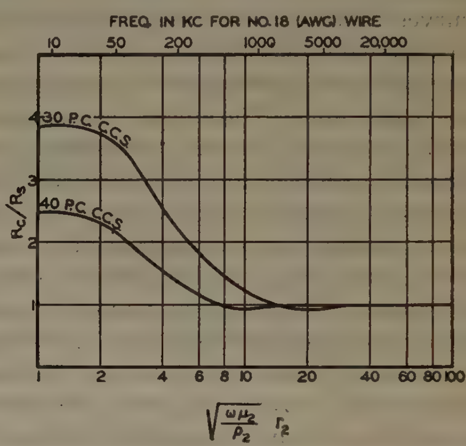


Fig. 4—Ratio of resistance of copper-covered steel to resistance of solid copper of same diameter at same frequency.

basis in which both copper and copper-covered-steel conductors have the same weight of copper per unit length. The copper-covered-steel conductor, because of its core, will necessarily have a greater diameter and a lower resistance. The advantage in resistance is small at low frequencies but is appreciable at radio frequencies.

For example, suppose that a No. 18 solid-copper wire has a satisfactory resistance. If the same weight of copper per unit length were used in a 30 per cent copper-covered-steel conductor, it would be approximately No. 12 in size and would have a resistance about half of that of the solid copper when the frequency is above 235 kilocycles. If the copper of the No. 18 conductor were employed in 40 per cent copper-covered steel, the size would be approximately No. 14 and the resistance about five eighths that of the solid conductor, provided the frequency is above 145 kilocycles. These frequency ranges represent limiting cases which show the most favorable comparison. Fig. 5 shows the resistance ratios of the two conductors with equal copper over the lower, less-favorable frequency range. The ordinates are the ratios of the alternating-current resistances of copper-covered steel to those of solid copper. Two scales of abscissas are shown: frequency when the solid conductor is No. 18, and a generalized scale which can be used with Fig. 3 for any size conductor.

Finally, the conductors are compared on the basis of equal resistance; that is, given that a solid conductor has a suitable resistance in a given frequency range, it is of interest to find the smallest size of copper-covered steel conductor which can be substituted without increasing the resistance.

This comparison, of course, depends on the frequency. At high enough frequencies the diameters must be the same if the resistances are to be equal; at lower frequencies the copper-covered steel must have a larger diameter. Fig. 6 shows for any frequency the diameter of copper-covered steel in terms of that of solid copper if the former is to have no higher resistance than the latter at all frequencies above the one considered. The curve thus gives a means of selecting a copper-covered-steel wire to replace a given copper wire that has been satisfactory, if the predominating factor underlying the choice of conductor is resistance.

It should also be observed that copper-covered steel has a tensile strength that may be as much as three

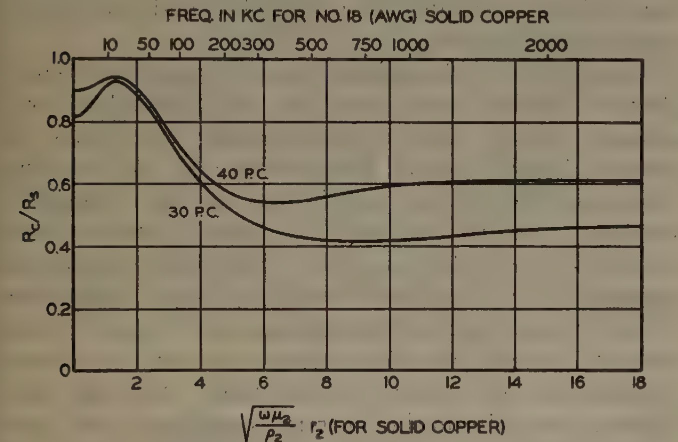


Fig. 5—Ratio of copper-covered steel to resistance of a solid-copper wire (of different size) which has the same weight of copper per unit length.

times that of solid copper for 30 per cent wire, the magnitude depending on the steel in the core, and may be as much as two and one half times for 40 per cent wire. Thus when mechanical strength is the principal factor determining the size of the conductor, as may well be the case in antenna installations, adequate strength can be obtained with a smaller size of copper-covered steel than of solid copper. Because of the smaller diameter, the copper-covered steel will have a higher resistance, but when mechanical strength and weight of the overhead construction are the determining factors, an increase of resistance may be permissible.

## Inductance of Copper-Covered Steel Wire

The inductance of a circuit has two components: one, attributable to flux existing entirely outside of the conductors, the other, usually much smaller than the first, attributable to flux within the conductors. The first component depends upon the geometric configuration of the circuit; that is, upon the location of the return paths, and also upon the diameter of the conductors. However, in the case of round wires it is independent of the materials of which they are composed and would

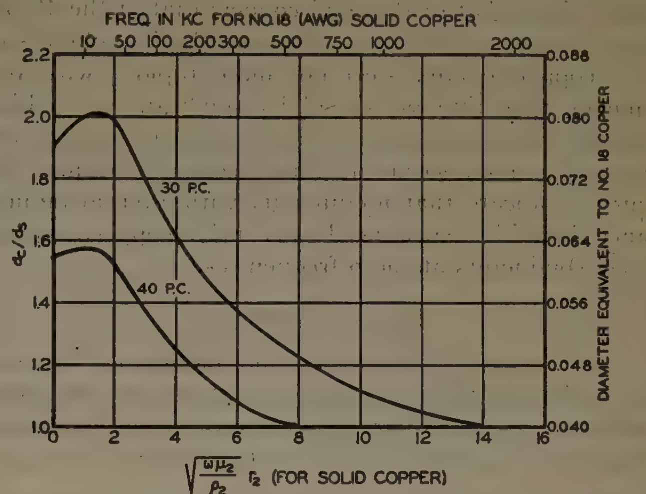


Fig. 6—Ratio of diameter of copper-covered steel to diameter of solid copper such that resistance is the same at and above the frequency given by abscissas. Upper and right scales apply to No. 18 wire.

be the same for both copper-covered steel and solid copper. The second component, the internal inductance, depends upon the materials of the conductor and upon their disposition. In the case of a solid cylindrical conductor of nonmagnetic material, this component is quite small, being $0.5\times10^{-7}$ henry per meter of length at zero frequency, and independent of diameter. The variation of internal inductance with frequency for a solid-copper conductor is shown in Fig. 7.

Since the internal inductance of a solid wire is proportional to the permeability, it might be expected that a copper-covered steel conductor, because of its steel core, would have an appreciably larger internal inductance than a solid-copper conductor. However, since

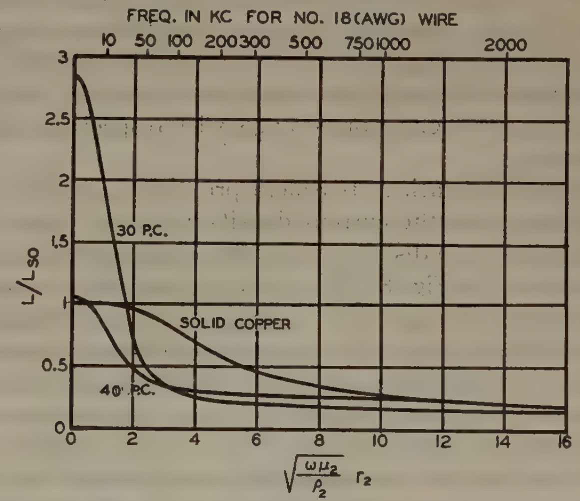


Fig. 7—Ratio of internal inductance to zero-frequency internal inductance of solid-copper wire.

only a small portion of the current flows in the core even at zero frequency, the effect of the steel is not proportional to core permeability. Fig. 7 shows the internal

inductance for both 30 and 40 per cent No. 18 copper-covered steel conductors. At zero frequency the inductance of the 30 per cent conductor is approximately three times that of the solid copper; that of the 40 per cent conductor, about equal to it. The inductances of the copper-covered steel fall more rapidly with frequency than that of the solid conductor, but at high enough frequencies, as in the case of resistance, the internal inductances of all three become the same. The curves indicate that no difficulty with internal inductance should be experienced when using copper-covered-steel conductors at radio frequencies.

## Physical Picture

The changes of resistance and inductance brought about by skin effect are readily explained by the redistribution of current in the conductor as the frequency increases. A current path near the center of the conductor has a greater flux linkage per unit of current than a path nearer the surface; hence the inductance of a path is greater the nearer it is to the center. The greater the inductance of a path, the less current it carries, and the greater is the lagging phase angle with respect to surface currents.

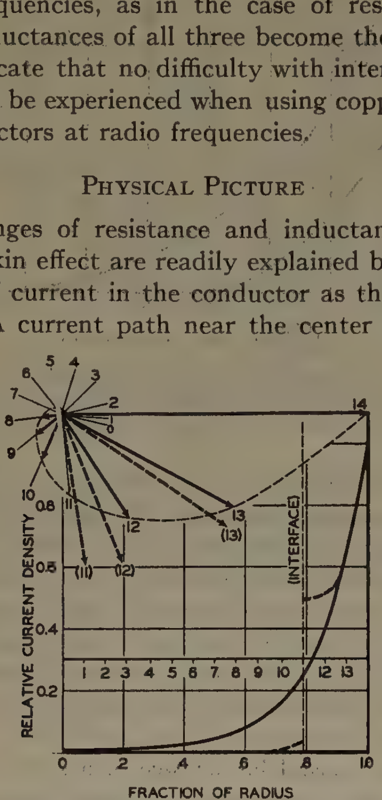


Fig. 8—Vector diagram and curve showing current densities at various locations for solid copper and 40 per cent copper-covered steel wires, when $r_2\sqrt{\omega\mu_2\rho_2}=9.8$.

The mathematical treatment provides a means of determining how the current density varies in phase and magnitude with distance from the center, and the result is depicted in Fig. 8. The upper portion shows a set of vectors which represent the magnitudes and phases of the current-density vectors at each of fourteen locations equally spaced along a radius. (Fourteen subdivisions for this particular ratio of $r_1/r_2$ made it possible to use the tables of Bessel functions without interpolation.) In the lower part of the figure the magnitudes of the vectors are shown as functions of radial distance. The dashed vectors and curve apply to the case of copper-covered steel; the solid vectors and curve, to a solid conductor of the same diameter and at the same frequency. Because of the interrelations between conductor size and frequency the curves apply to No. 24 wire at 3250 kilocycles, No. 18 at 840 kilocycles, No. 12 at 210 kilocycles, or No. 6 at 54 kilocycles, to name a few of the possible combinations.

The current-density vectors and curve were calculated for the corresponding tube and are almost exactly the same as those in the copper part of the copper-covered steel. They are omitted from Fig. 8 for the sake of clarity.

The vectors for the solid conductor of the figure indicate a phase change from surface to center of slightly more than 360 degrees. The current densities with phases near 180 degrees indicate current whose instantaneous directions are opposite to the surface currents. These reversed currents produce losses and also subtract from the total current, thus making for a large effective resistance, or ratio of $i^2R$ loss to square of total current. Removing the material from the center of the conductor eliminates the reversed currents and reduces the resistance; therefore, it would be expected that a tube of suitable inner diameter at this frequency would have a lower resistance than the solid conductor of the same outer diameter.

In addition to the elimination of reversed currents, there is another reason why the tube has a lower resistance; namely, the more favorable distribution of current density. At the inner surface of the tube the magnetic field is zero and it may be shown mathematically that this requires the slope of the current-density curve to be zero there. This raises the curve appreciably over that of the solid conductor, making for a better utilization of the material and a lower resistance. In general, the more nearly uniform the current distribution, the lower the resistance. For the case shown, the alternating-current resistance per unit length of the tube is 92 per cent of that for the solid conductor. This corresponds to $br_2=9.8$ for the 40 per cent case in Figs. 2 and 4.

The resistance and inductance of copper-covered steel at all but the lowest frequencies are the same as those of the corresponding copper tube. This is true because the steel core carries negligible current as has been determined from the current-density equations derived in conjunction with the resistance equation. Thus as far as electrical behavior is concerned at radio frequencies, copper-covered steel may be regarded as a tubular conductor.

## Skin Thickness

As the frequency is increased to higher and higher values, the current density falls off more and more rapidly with distance in from the surface. Mathematically, the Bessel functions, in terms of which the current density is expressed, approach similarity to decreasing exponential functions; and if the frequency becomes high enough, the current-distribution curves for tubular and copper-covered steel conductors become the same as those for the solid one; that is, in all three the magnitude of current density decreases

exponentially to zero. This limiting case corresponds to that discussed by Rayleigh.[5] The decreasing exponential is characterized by an attenuation constant, and since the current density is alternating and the phase changes linearly with distance, there is also a phase constant. Expressed respectively as the distance required for the magnitude to fall to $1/e$ of its value at the surface, and as the distance for the phase of the current density to change by 1 radian, the two constants are equal and given by $b/\sqrt{2}$. The reciprocal $\sqrt{2}/b$ is the "skin thickness" or depth of penetration[6,7] and for copper at 20 degrees centigrade has the value $2.64/\sqrt{f}$ inches. At a depth equal to three skin thicknesses, the current density has a value 5 per cent of that at the surface and lags by 3 radians or nearly 180 degrees.

In the limiting high-frequency case, the total root-mean-square current, which is the integral of current density over the cross section with due regard to phase, approaches in magnitude the product $J_s\delta p/\sqrt{2}$, where $J_s$ is the root-mean-square magnitude of surface current density, $p$ the perimeter, and $\delta$ the skin thickness, all in m.k.s. units. As the frequency increases without limit, the phase of the total current approaches an angle of 45 degrees lagging behind the phase of the surface current density. This phase relationship of 45 degrees means that the resistance per unit length approaches equality with the internal reactance per unit length. This equality together with the information in Fig. 1 permits a rapid determination of the internal inductance of copper-covered steel when the frequencies are high enough so that the limiting case holds.

The resistance per unit length of the conductor in the limiting range is simply the quotient of resistivity by skin area or $\rho/\delta p$; that is, the resistance is the direct-current resistance of the "skin." Thus the concept of skin thickness may be employed to determine the required size of conductor such that the conductor has a given resistance at any specified frequency in the limiting range. When $br_2$ is 30 or more, the error is 2 per cent or less.

Also, the ratio of the depth of copper in copper-covered steel to skin thickness is a measure of whether there is enough copper to make the greatest possible use of copper to increase the conductance. From Fig. 4 it follows that if $br_2$ is 14 for the 30 per cent conductor or 8 for the 40 per cent conductor which, respectively, correspond to a copper depth of 1.3 and 1.1 skin thicknesses, the resistance of the copper-covered steel is at least as low as that of a solid-copper conductor of the same diameter. In Fig. 8 the copper depth is 1.4 skin thicknesses.

[5] Lord Rayleigh, "Scientific Papers," volume II, Cambridge University Press, Cambridge, England, 1900, page 486.

[6] C. P. Steinmetz, "Transient Electric Phenomena and Oscillations," McGraw-Hill Book Company, New York, New York, 1909.

[7] Harold A. Wheeler, "Formulas for the skin effect," Proc. I.R.E., vol. 30, pp. 412–424; September, 1942. This paper also includes an excellent list of references on skin effect.


### Acknowledgment

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# Corrective Networks for Feedback Circuits*


VINCENT LEARNED†, ASSOCIATE, I.R.E.



*Summary*—**Design information is given for simplified corrective networks for application with negative-feedback devices. The corrective networks are applied to control the cutoff attenuation characteristic of the feedback transmission loop to prevent oscillation. To provide a design factor of safety against oscillation, the cutoff transmission characteristic should not attenuate at a rate of more than 10 decibels per octave. Corrective networks are required to achieve this attenuation rate.**


## Introduction

The requirements for stabilizing amplifying devices with negative feedback against oscillation have been given in the technical literature in recent years.[1–4] These requirements relate to certain necessary conditions of phase shift and attenuation in the amplifier and feedback circuits. To obtain an optimum design that satisfies these requirements, it has been shown in previous literature that definite attenuation characteristics must be followed. Networks are given in this article which may be used to obtain amplifying devices that will approach the optimum characteristics. These networks consist of simple circuit-element combinations that are suited for operation in the plate and grid circuits of vacuum-tube amplifiers.

Fig. 1 shows a schematic feedback amplifier with a gain of $\mu\underline{/\theta}$ and a feedback network of $\beta\underline{/\psi}$. The net gain of the amplifier is given by the relation

$$\text{net gain} = \mu\underline{/\theta}\,/(1-\mu\underline{/\theta}\,\beta\underline{/\psi}). \quad (1)$$

In the normal frequency range the phase of $\mu\underline{/\theta}\ \beta\underline{/\psi}$ is adjusted to give a feedback action that opposes the applied signal, thereby reducing the gain and giving negative feedback. The quantity $\mu\underline{/\theta}\ \beta\underline{/\psi}$ is the


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† Sperry Gyroscope Company, Inc., Garden City, L. I., New York.


[1] H. Nyquist, "Regeneration theory," *Bell Sys. Tech. Jour.*, vol. 11, pp. 126–147; January, 1932.

[2] H. W. Bode, "Relations between attenuation and phase in feedback amplifier design," *Bell Sys. Tech. Jour.*, vol. 19, pp. 421–454; July, 1940.

[3] F. E. Terman, "Network theory, filters, and equalizers—Part II," Proc. I.R.E., vol. 31, pp. 235–240; May, 1943.

[4] H. S. Black, "Stabilized feedback amplifiers," *Elec. Eng.*, vol. 53, pp. 114–120; January, 1934.

transmission characteristic of the "feedback loop" and may be measured by breaking the circuit at $X$ of Fig. 1 and applying a signal at $A$ and comparing it with the signal obtained at $B$. According to the Nyquist stability criteria the system will not oscillate if the transmission characteristic $\mu/\theta\ \beta/\psi$ has a gain less than unity when the phase shift reaches 180 degrees.

In the normal frequency range of a feedback device, $\mu/\theta\ \beta/\psi$ is usually nearly constant. Outside the normal

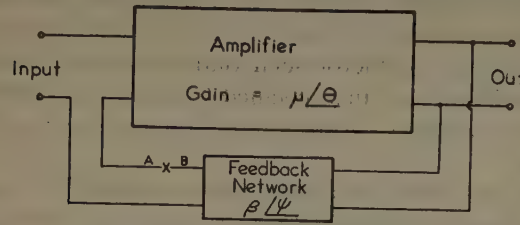


Fig. 1—Schematic of basic feedback amplifier.

frequency range it is desirable to attenuate as rapidly as possible to reduce the gain to less than unity. The only thing that limits a rapid reduction in gain is the phase shift that occurs with the change in gain. The attenuation characteristic outside the normal frequency band must be controlled to prevent the phase shift from exceeding the 180-degree maximum.

## Characteristics of Amplifiers

The cutoff characteristic of an amplifier outside of its normal frequency range is determined by the reactive elements of the interstage coupling devices employed. In a resistance-coupled amplifier these usually are stray circuit capacitances and grid-coupling condensers. Each amplifier stage introduces reactive elements which combine to produce a cutoff characteristic that may or may not produce a stable feedback amplifier. The cutoff characteristic usually must be modified to obtain a stable amplifier.

Bode[2] has shown for minimum phase-shift networks that the phase-shift characteristic of a network may be obtained directly from its amplitude characteristic. Thus in the usual amplifier circuits the phase-shift characteristic is directly related to the amplitude characteristic and is independent of the particular circuit combination which produced it. To obtain a special phase-shift characteristic for a feedback amplifier, the corresponding amplitude response needs to be considered.

The usual coupling network with reactive elements has a frequency region for which the amplitude response is uniform as well as a region for which the response varies with frequency. There are certain universal facts associated with these coupling devices operating well into the cutoff region to give an asymptotic response.

1) The amplitude-frequency response is directly or inversely proportional to frequency as the case may be.
2) The phase shift is 90 degrees lagging for the response decreasing with frequency and 90 degrees leading for the response increasing with frequency.
3) With logarithmic co-ordinates the slope is 1 or −1.
4) With logarithmic frequency and decibel amplitude scale the slope is approximately 6 decibels per octave.
5) No network in the asymptotic cutoff region will have a slope of less than unity and all more complicated networks have asymptotic slopes that are integers when plotted on logarithmic co-ordinates.
6) The phase shift is corresponding integral multiples of 90 degrees.

## Feedback Requirements

The greatest cutoff attenuation rate that may be employed over any extended frequency spectrum in a feedback device is 12 decibels per octave. This corresponds to a phase shift of 180 degrees which is the limit for stability. In a practical device a margin of safety must be provided, so that a slope of less than 12 decibels per octave must be used. It is customary to allow a 30-degree phase-shift margin, thus giving a slope of $150/180 \times 12 = 10$ decibels per octave.

The usual amplifier has an asymptotic cutoff characteristic which is greater than 10 decibels per octave since each coupling circuit produces 6 decibels per octave or some multiple of 6 decibels per octave. The design of a stable feedback amplifier requires a transition from the region of controlled 10-decibel-per-octave attenuation in which the feedback loop gain is greater than unity,

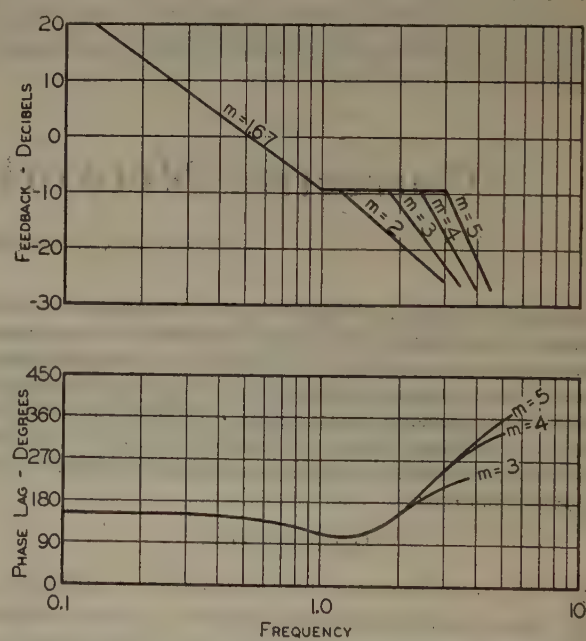


Fig. 2—Ideal amplitude cutoff response with resulting phase shift. $m$ = slope in units of 6 decibels per octave. $m = 1.67$ is controlled 10-decibel-per-octave portion of cutoff. $m = 2, 3, 4$, or 5 is uncontrolled asymptotic cutoff.

to the region of uncontrolled attenuation in which the attenuation rate may be very high and the loop gain is much less than unity.

This transition may be obtained by following Bode's ideal cutoff transmission characteristic for the "feedback loop" as shown in Fig. 2. The slope of the various

segments of the curve are given in units of $m$ which are units of 6 decibels per octave. This cutoff characteristic satisfies Nyquist's stability criteria, since the phase shift is less than 180 degrees for frequencies where the transmission characteristic has a gain greater than unity.

This attenuation curve features a factor of safety against changes in both gain and phase shift. The curve attenuates at a rate of 10 decibels per octave ($m=1.67$), which provides a phase margin of safety of 30 degrees with respect to the 180-degree maximum.

In addition, a region of zero-gain change is provided to give a phase-shift cancellation effect against the larger phase shift produced by the rapid asymptotic cutoff at the higher frequencies. The frequency range of this step in the attenuation characteristic is equal to the ratio of the slope of the asymptotic characteristic to the slope of the controlled characteristic. The attenuation step is 9 decibels below the unit loop gain (zero-decibel) line, giving a factor of safety against gain variations in the amplifier. The curves illustrated in Fig. 2 apply to the high-frequency cutoff characteristic while similar curves apply equally well to the low-frequency cutoff characteristic.

## Modification of the Feedback-Loop Amplitude Response

The cutoff amplitude-response characteristic of most feedback devices will not satisfy the slope requirements

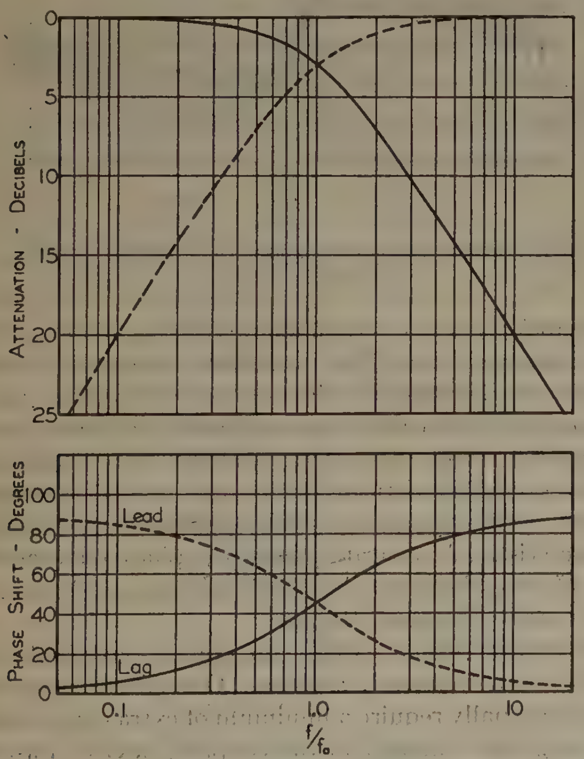


Fig. 3—Transmission characteristic of resistive-reactive coupling elements. $X=R$ at $f/f_0=1$.

of the ideal cutoff characteristic. The usual response characteristic cuts off too rapidly or allows no margin of safety. Corrective networks must be added to control the attenuation rate and to provide the necessary step in the attenuation characteristic.

All simple coupling combinations conventionally used with vacuum tubes provide an asymptotic response

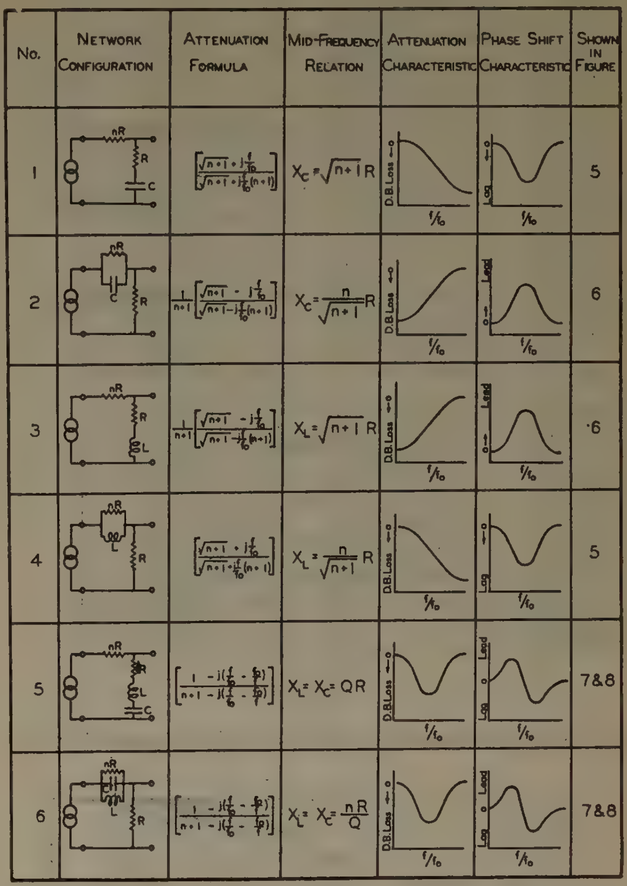


Fig. 4—Design data for corrective networks. Generator is considered to have zero impedance. Load impedance is considered infinite.

characteristic that is a multiple of 6 decibels per octave. For example, the low-frequency response of a transformer-coupled stage will cut off at an asymptotic rate of 6 decibels per octave caused by the shunting effect of the inductance upon the resistive parts of the circuit. At the high frequencies the asymptotic cutoff may be at a rate of 12 decibels per octave due to the combination of leakage inductance and shunt capacitance. Likewise, each stage of resistance coupling gives a 6-decibel-per-octave asymptotic cutoff at the high and low frequencies. Fig. 3 is a plot of the ordinary low- and high-frequency cutoff characteristic for resistance-reactance elements.

To obtain a transmission-slope characteristic of 10 decibels per octave, which is not a multiple of 6 decibels, it is necessary to include corrective networks. To give a net 10-decibel-per-octave slope the corective network may contribute a 4-decibel-per-octave slope to be added to a 6-decibel-per-octave slope, or it may subtract a 2-decibel-per-octave slope from a 12-decibel-per-octave slope. Corrective networks 1, 2, 3, and 4 of Fig. 4 illustrate four different combinations of circuit elements that will give a slope of less than 6 decibels per octave over a limited frequency range. Networks 5 and 6 of

Fig. 4 show two different combinations which may be used to obtain a step in the frequency-response char-

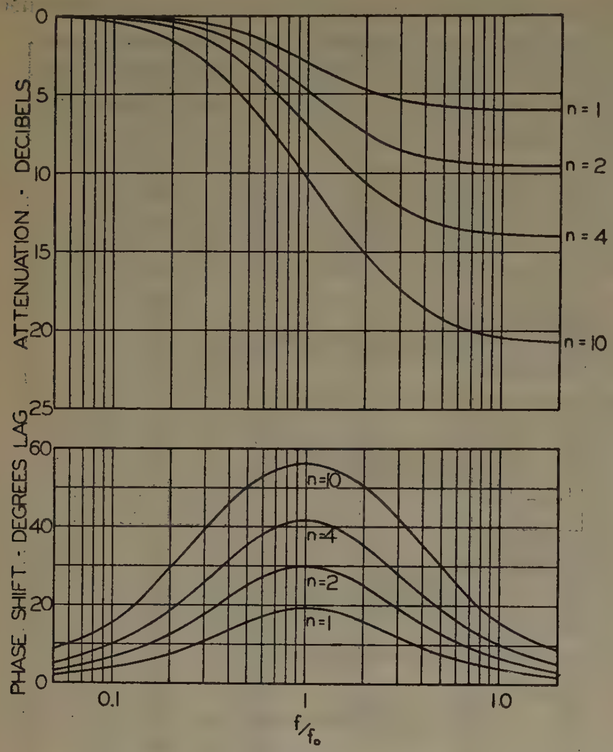


Fig. 5—Response characteristics for corrective networks 1 and 4.

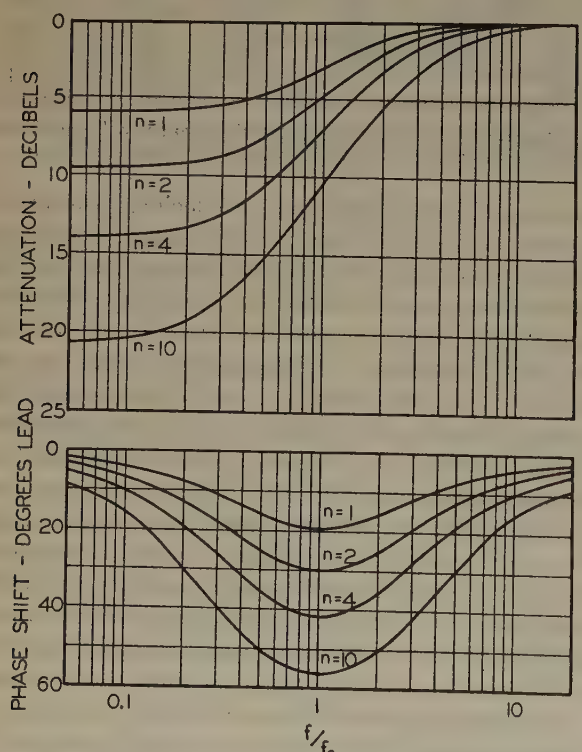


Fig. 6—Response characteristics for corrective networks 2 and 3.

acteristic. The response characteristics of these networks are shown in Figs. 5 to 8.

Each corrective network contains two resistance elements which are related in magnitude by a parameter $n$. The reactance of the associated capacitance or inductance is obtained by an expression derived for the mid-frequency of the network. Networks 5 and 6 containing both inductance and capacitance are related to the resistive components at the mid-frequency (resonance) by a factor $Q$. The mid-frequency relation for each corrective network is given in Fig. 4.

## Application of Networks

The systematic design of a feedback amplifier using these corrective networks requires the determination of the attenuation or gain that must be introduced. This may be obtained by first plotting in decibels the desired

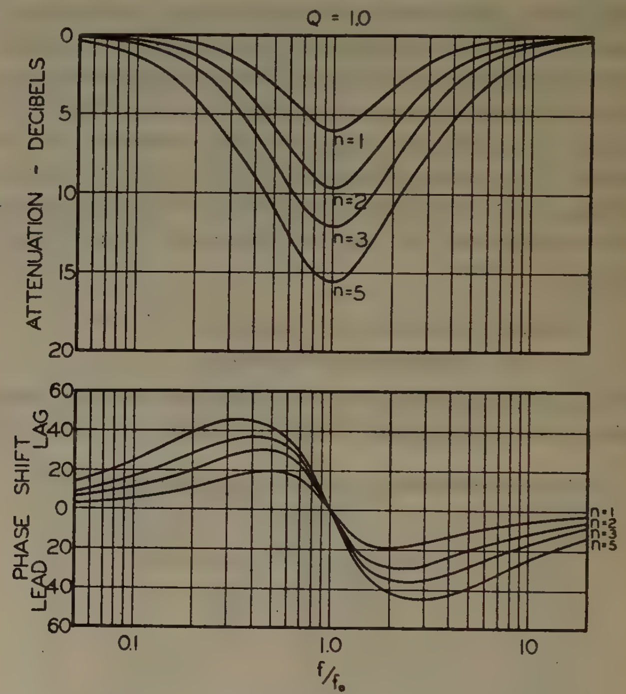


Fig. 7—Response characteristics for corrective networks 5 and 6, $Q=1.0$.

10-decibel-per-octave cutoff characteristic on a logarithmic frequency scale. The amplitude responses of each stage of the amplifier are then plotted either from measurements or from calculations and using the accompanying curves. With these data available the amplitude-response characteristic of each stage may be modified to give a total response which approaches the desired characteristic. A separate plot of the phase characteristics offers a check on the amplitude plot and may be used to plot a Nyquist diagram.

There are several circuit possibilities available for each desired frequency response. One circuit combination will usually require a minimum of extra components by using those already in the coupling device. Two resistive circuit components are inherently available: the grid-return resistor, and the equivalent-generator impedance. These may both be used to advantage in applying the networks of Figs. 4 and 9. Typical vacuum-tube-circuit arrangements with their equivalent circuits are shown in Fig. 9. Once the corrective network circuit

configuration is decided upon and its mid-frequency is determined from an amplitude-response plot, sufficient data are available to complete the network design.

Approximations are often useful in simplifying network design. For example, the equivalent-generator impedance is often conveniently considered negligible, or two corrective networks may be used in tandem with

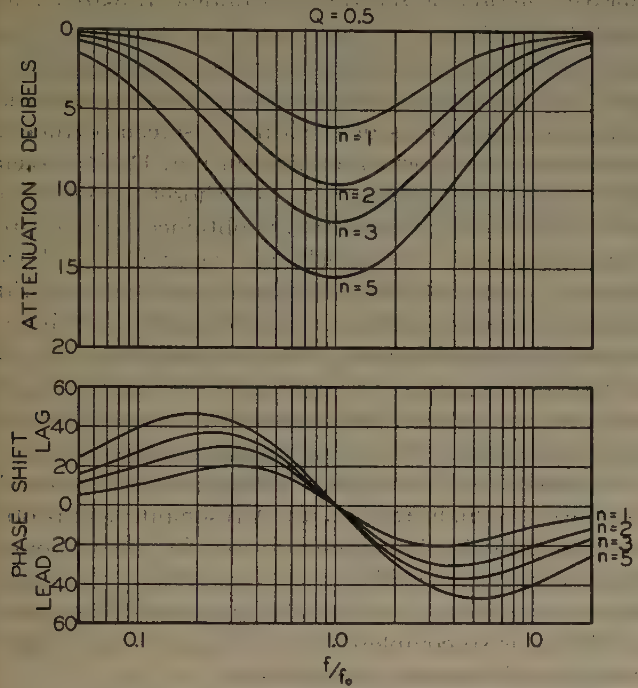


Fig. 8—Response characteristics for corrective networks 5 and 6, $Q=0.5$.

the impedance of the first considered negligible with respect to the impedance of the second. Circuit simplification is often possible. For example, a given circuit element may be useful in one network configuration for attenuating at low frequencies and may be useful in another configuration for attenuation at the high frequencies. It is possible to employ one network attenuating over a portion of the frequency spectrum and another network picking up where the first stops. The

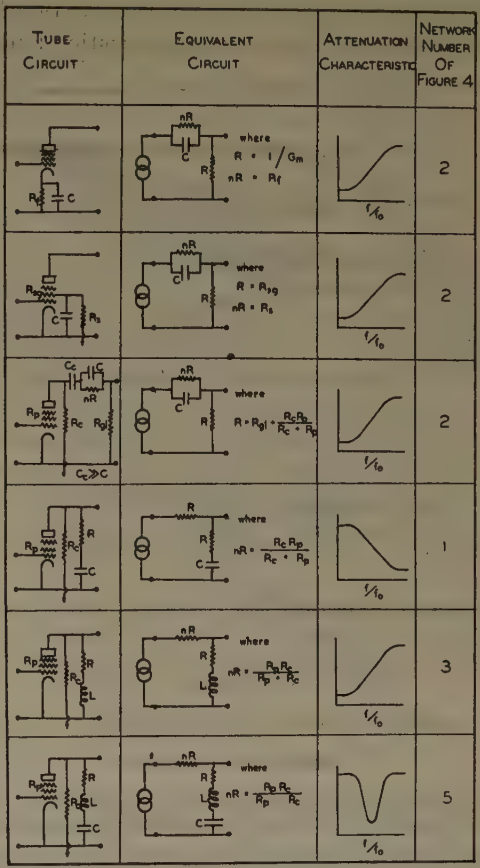


Fig. 9—Typical practical vacuum-tube circuits illustrating use of corrective networks.

slope contributed by these networks is governed by two factors: the parameter $n$, and the frequency separation between the mid-frequency points of each network.

Networks 5 and 6 of Fig. 4 are used in the same

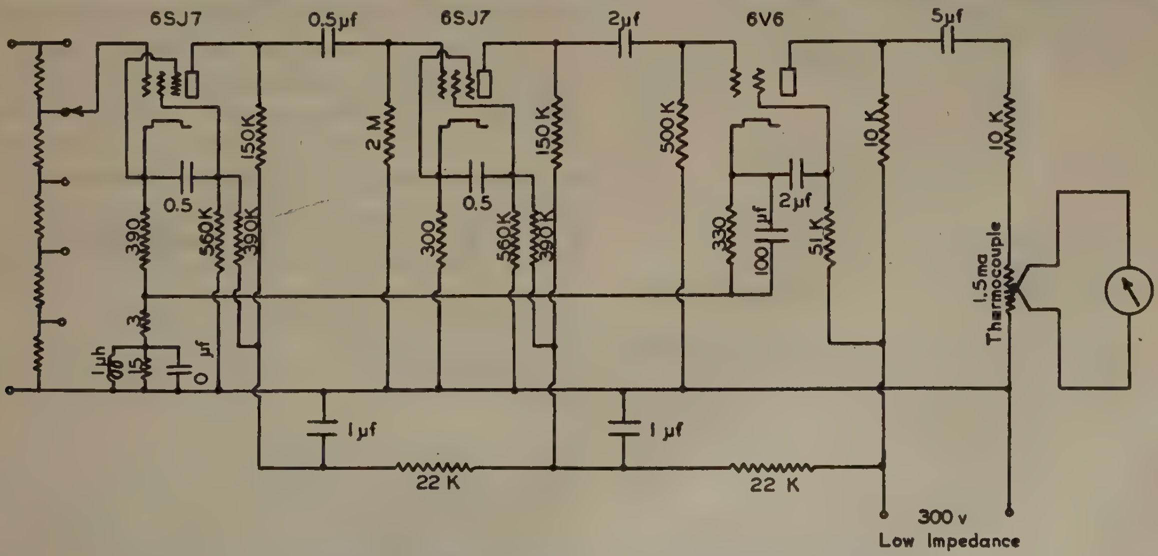


Fig. 10—Circuit diagram of feedback amplifier stabilized by use of corrective networks. $R_{sg}$ for 6SJ7 = 75,000 ohms, $n=3.1$. $R_{sg}$ for 6V6 = 21,000 ohms, $n=2.4$.

manner as the others. They provide a response characteristic which first attenuates and then releases. They are useful in producing the attenuation step of the ideal cutoff characteristic of Fig. 2. The desired step is accomplished by combining the response of these networks with a uniform 6-decibel-per-octave characteristic. The phase-

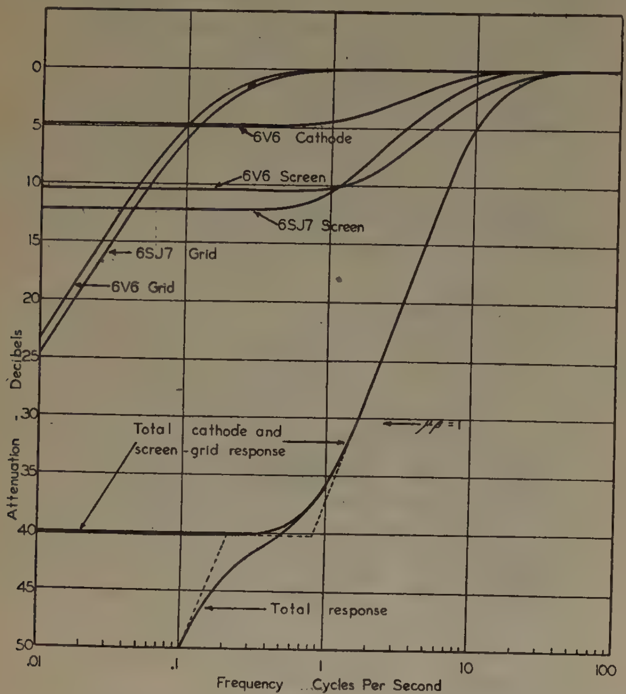


Fig. 11—Plot of low-frequency response for amplifier of Fig. 10. Response is controlled by screen and cathode degeneration. The response of each screen and cathode is added to give step characteristic. Interstage coupling circuits cutoff at lower frequency.

response characteristic for these networks shows a phase-canceling property which helps to avoid oscillation. The selection of the parameters $n$ and $Q$ makes the networks adaptable to various circumstances. There are other resonant configurations for which these curves may be used for approximating the response.

## Example

Fig. 10 is the circuit diagram of a feedback amplifier stabilized with the use of corrective network. Negative-current feedback is obtained by producing a voltage in the cathode circuit of the first stage that is proportional to the current in the third stage.

The low-frequency response is corrected by the degenerative action of the screen-dropping resistors and the self-bias resistors. This degeneration occurs when the reactance of the by-pass condensers becomes large compared to the circuit resistance as the frequency is lowered. The responses of the various screen, cathode, and interstage networks are given in Fig. 11. It is seen that if the circuit parameters are chosen properly a region of controlled attenuation is obtained with a rate of approximately 10 decibels per octave. The low-frequency cutoff of the interstage-plate control-grid coupling networks is adjusted to give a step in conjunction with the low-frequency flattening off of the screen and cathode responses. The equivalent circuits of Fig. 9 were used to compute the screen and cathode circuit parameters to give the desired response characteristic.

The resonant circuit in the feedback path is employed to stabilize the high frequencies. A resonant peak is obtained to provide the necessary step in the response characteristic which aids in the cancellation of phase shift in the region of the oscillation point. The phase cancellation is so complete that no regeneration occurs in the over-all response characteristic.

The use of the type of corrective networks shown in this paper offers an approximate method for designing stable feedback devices. Network simplicity is sacrificed for exact response characteristics which is sufficient for many applications. The method of stabilizing feedback devices by obtaining a prescribed amplitude response characteristic is due to H. W. Bode of the American Telephone and Telegraph Company, and an understanding of the results of his work on this subject is quite essential to the proper design of feedback devices.

# Frequency Modulation of Resistance-Capacitance Oscillators*

MAURICE ARTZT†, ASSOCIATE, I.R.E.

*Summary*—**A method is described for direct-frequency modulation of a resistance-capacitance oscillator which is simpler and more stable than the beat oscillators formerly used. Spurious amplitude modulation is reduced to a negligible value without the use of limiters and without introducing appreciable harmonics in the output wave. Balanced control tubes prevent transients or signal frequencies from appearing in the output. Curves are given which provide an easy method of choosing the network and constants for any desired condition. The device is especially suited to facsimile transmission by subcarrier frequency modulation.**

## Introduction

THE USES of frequency-modulated subcarriers in facsimile and communications systems have greatly multiplied in the past few years.[1,2] In all of these instances the frequency swing is a sizable percentage of the carrier frequency, and cannot be obtained directly by reactance-tube control of an inductance-capacitance tank-circuit oscillator. The usual method has been to use reactance tube control of one or both tanks in a beat oscillator, thus obtaining a low-frequency carrier with the same frequency shift as the controlled oscillator.

The frequency-modulated resistance-capacitance oscillator[3] here described replaces the beat-oscillator system and has been found to be far more stable and considerably simpler in adjustment. The large frequency swings are obtained directly without heterodyning and the frequency drift usually associated with beat oscillators is thus avoided.

The frequency of oscillation of any resistance-capacitance oscillator is determined by the constants of the network, and therefore, changing any resistance or capacitance value will change the frequency. Replacing any of the resistive elements with a tube allows the frequency to be controlled by this tube.

There are two general classifications of resistance-capacitance oscillators, those having zero phase shift in the network,[4–6] and those having 180-degree phase shift in the network.[7] Tubes may be used to replace one or more of the resistors in oscillators of either type, though there are definite advantages in using oscillators with 180-degree phase-shift ladder networks. In either case, changing any single element of the network generally will change the network loss and give some undesired amplitude modulation in addition to the frequency modulation. These amplitude variations are more easily eliminated in the ladder network oscillator without resorting to automatic volume controls which limit the speed of response of the system. By proper choice of constants and operating conditions the amplitude modulation can be reduced to a negligible amount even with shifts as high as $\pm 40$ per cent of the carrier frequency, and without introducing appreciable harmonic distortion.

The particular frequency-modulated oscillator used in one facsimile system[8] will be described first. Later, more general data on design will be given.

## The Modulated Oscillator

The basic circuit of the oscillator, Fig. 1, is a four-step series-capacitor ladder network driven by a pentode $T_1$ and cathode follower $T_2$. If all network resistors were alike a voltage gain of 18.36 would be required to overcome the network loss and cause oscillation.[9] But in this case $R_2$ is made enough smaller than the other three re-

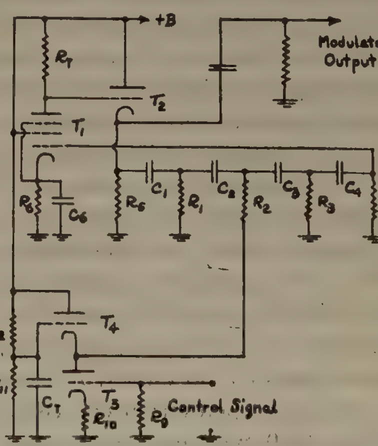


Fig. 1—Frequency-modulated resistance-capacitance oscillator.

sistors, $R_1$, $R_3$, and $R_4$, to make the network loss when $R_2$ is grounded about equal to the loss when $R_2$ is open-circuited. With $R_2$ open the fundamental frequency is 2000 cycles, and with $R_2$ grounded it is somewhat over 4000 cycles. Values of resistance in series with $R_2$

* Decimal classification: R355.9×R414. Original manuscript received by the Institute, December 15, 1943.

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[1] R. E. Mathes and J. N. Whitaker, "Radio facsimile by subcarrier frequency modulation," *RCA Rev.*, vol. 4, pp. 131–154; October, 1939.

[2] Warren H. Bliss, "Use of subcarrier frequency modulation in communication systems," Proc. I.R.E., vol. 31, pp. 419–423; August, 1943.

[3] U. S. Patent 2,321,269, June 1943.

[4] C. K. Chang, "A frequency-modulated resistance-capacitance oscillator," Proc. I.R.E., vol. 31, pp. 22–25; January, 1943.

[5] H. H. Scott, "A new type of selective circuit and some applications," Proc. I.R.E., vol. 26, pp. 226–236; February, 1938.

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[7] E. L. Ginzton and L. M. Hollingsworth, "Phase-shift oscillators," Proc. I.R.E., vol. 29, pp. 43–49; February, 1941.

[8] U. S. Patent 2,326,740; August, 1943.

[9] This ratio is obtained from the equations of the network in Fig. 3A.

intermediate between zero and infinity will give frequencies between these limits. These varying values of resistance are furnished by the tube circuit $T_3$ and $T_4$.

The control circuit has the direct-current component of its output balanced out, so that changing the equivalent resistance in series with $R_2$ does not introduce into the network transients or modulating signal frequencies. The resistors $R_{11}$ and $R_{12}$ form a center tap on the B voltage, and $T_4$ is connected as the plate resistor of $T_3$. If $T_3$ is at zero bias, it draws full plate current

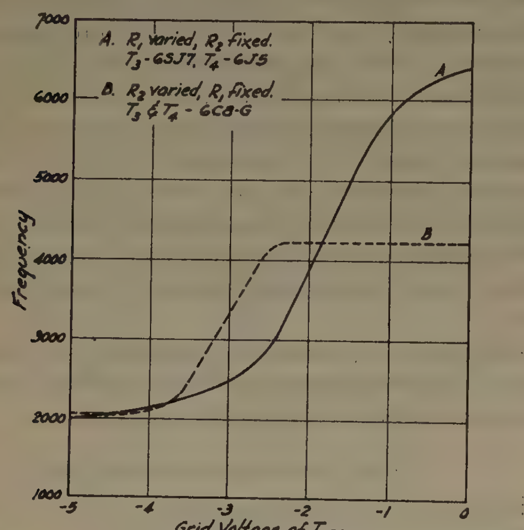


Fig. 2—Frequency-modulation characteristics of oscillator in Fig. 1.

through $T_4$, and the two tubes together form a low-resistance path from $R_2$ to ground. Thus $R_2$ is effectively grounded for the oscillation frequency, while at a direct-current potential of half the B voltage, and a four-step ladder with a frequency slightly above 4000 cycles is obtained. Now as negative signal is applied to $T_3$ grid, its equivalent plate resistance is increased. The plate voltage applied to $T_3$ cannot rise appreciably because the cathode-follower action of $T_4$ makes it draw less plate current at the same time, and increases its plate resistance. $T_3$ and $T_4$ then present a higher equivalent resistance in series with $R_2$, but at the same direct-current potential above ground. This lowers the network frequency. In the extreme case of $T_3$ at cutoff, $T_4$ is also cut off, and the return of $R_2$ is open-circuited, thus bringing the network to the 2000-cycle end of its range.

This control system can be used to vary any of the resistors in the network. The curves in Fig. 2 illustrate two conditions. For curve $A$ the resistor $R_1$ was varied by a tube combination of a 6SJ7 for $T_3$ and a 6J5 for $T_4$. $R_2$, $R_3$, and $R_4$ were equal and all grounded. For curve $B$, resistor $R_2$ was varied and $R_1$ fixed. The control tubes in this second case were the two triodes of a 6C8-G tube.

The linearity of the modulation will be determined by which resistor or resistors are varied, and by the characteristic of $T_3$. Methods of making the linear range very wide are described later. In this case curve $A$ is linear from 3000 to 5800 cycles, a swing of $\pm 32$ per cent on a carrier of 4400 cycles. Curve $B$ is linear from 2300 to 4000 cycles, or a swing of $\pm 27$ per cent on a carrier of 3150 cycles.

In both cases the amplitude modulation is negligible over the linear range, and only a few per cent at the extreme ends of swing. This would not be true, however, except for the precautions described under "Amplitude Characteristics."

The speed of operation is apparently instantaneous. Oscilloscope observations have shown, for instance, that when using the system as for curve $B$, the oscillator can be shifted from 3000 to 4000 cycles, held on 4000 cycles for only 1 cycle, and then shifted back to 3000 cycles without any transients observed in the signal either before or after demodulation in a suitable discriminator. Such precise following of a square-wave signal makes the method very useful for facsimile transmission.

## Frequency Ranges

### A. Four-Step Ladder Network

The frequency range to be expected and the best operating conditions can be determined from calculated curves of the networks. Fig. 3A shows the four-step ladder as used in Fig. 1. All capacitors are assumed equal,

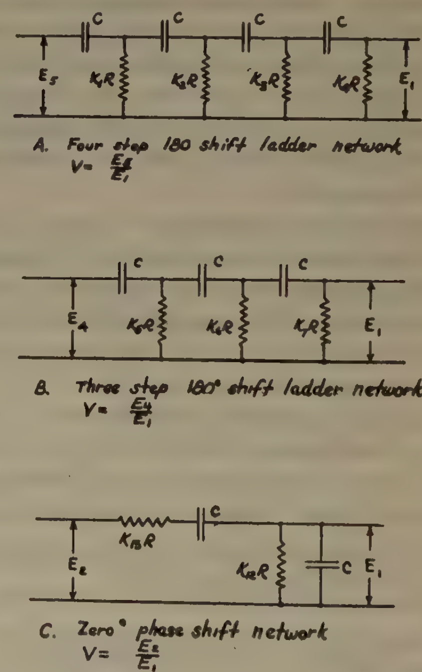


Fig. 3—Three types of resistance-capacitance networks.

and for simplicity in calculations $R$ is assumed to be 1. Thus the complex equations for $E_5/E_1$ have only the variables $X_c$ and $K$. Frequency is plotted as the numerical value of $1/X_c$, and to obtain actual cycles per second the $F$ ordinate must be multiplied by $1/2\pi RC$. Under these conditions the equation for $E_5/E_1$ is

$$\frac{E_5}{E_1} = V = \frac{K_1K_2K_3K_4 - X^2(3K_1K_2 + 4K_1K_3 + 2K_1K_4 + 3K_2K_3 + 2K_2K_4 + K_3K_4) + X^4}{K_1K_2K_3K_4} + j\left[\frac{X^3(2K_1 + 2K_2 + 2K_3 + K_4) - X(4K_1K_2K_3 + 3K_1K_2K_4 + 2K_1K_3K_4 + K_2K_3K_4)}{K_1K_2K_3K_4}\right]. \quad (1)$$

If the first resistor only is varied, then $K_2 = K_3 = K_4 = 1$,

$$\text{therefore } V_1 = \left[\frac{K_1 - X^2(9K_1 + 6) + X^4}{K_1}\right] + j\left[\frac{X^3(2K_1 + 5) - X(9K_1 + 1)}{K_1}\right]. \quad (2)$$

When oscillating, the network shift is 180 degrees, and the ($j$) term, therefore, must be zero.

$$\text{Therefore} \quad X^2 = (9K_1 + 1)/(2K_1 + 5) \quad (3)$$

$$\text{and} \quad F_1 = 1/X = \sqrt{(2K_1 + 5)/(9K_1 + 1)}. \quad (4)$$

This is plotted as $F_1$ in Fig. 4, and shows the relative

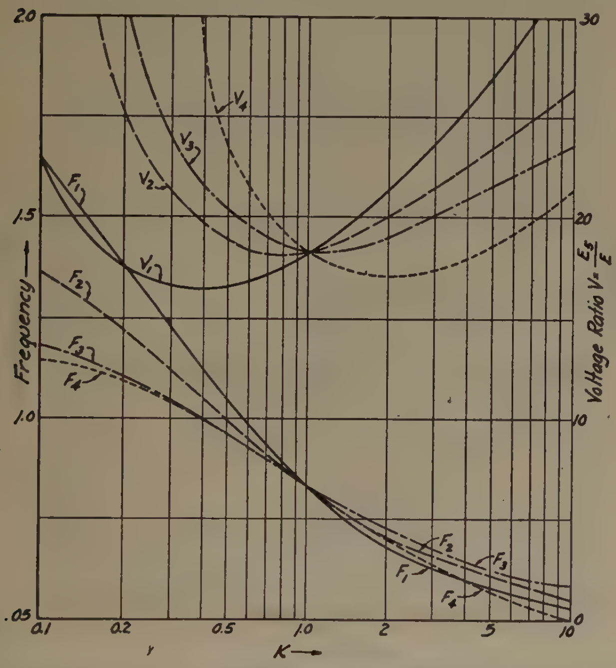


Fig. 4—Frequency and voltage ratio characteristics of four-step ladder network in Fig. 3A.

frequency for which the network phase shift is 180 degrees as $K$ is varied.

By placing the value of $X^2$ from (3) in (2), a plot is obtained of the voltage ratio of input to output ($V_1$) as $K$ varies. This is also plotted in Fig. 4 as $V_1$. Actually it is a negative number, but plotted positive here for convenience.

In a similar manner, if $R_2$ only is varied, $K_1 = K_3 = K_4 = 1$, and a new equation for $V_2$ is obtained. From this the curves $F_2$ and $V_2$ in Fig. 4 are plotted $F_3$, $V_3$ and $F_4$, $V_4$ are similarly derived from (1) by using either $K_3$ or $K_4$ as the variable.

*B. Three-Step Ladder Network*

The ladder network may be used with only three steps if desired, as in Fig. 3B. With this network, and similar conditions of $R = 1$ and all capacitances equal, the equation for $V = E_4/E_1$ becomes

$$V = \left[\frac{K_5K_6K_7 - X^2(2K_5 + 2K_6 + K_7)}{K_5K_6K_7}\right] + j\left[\frac{X^3 - X(3K_5K_6 + 2K_5K_7 + K_6K_7)}{K_5K_6K_7}\right]. \quad (5)$$

As before, to vary the first resistor only make $K_6 = K_7 = 1$, then

$$V_5 = \left[\frac{K_5 - X^2(3 + 2K_5)}{K_5}\right] + j\left[\frac{X^3 - X(5K_5 + 1}{K_5}\right]. \quad (6)$$

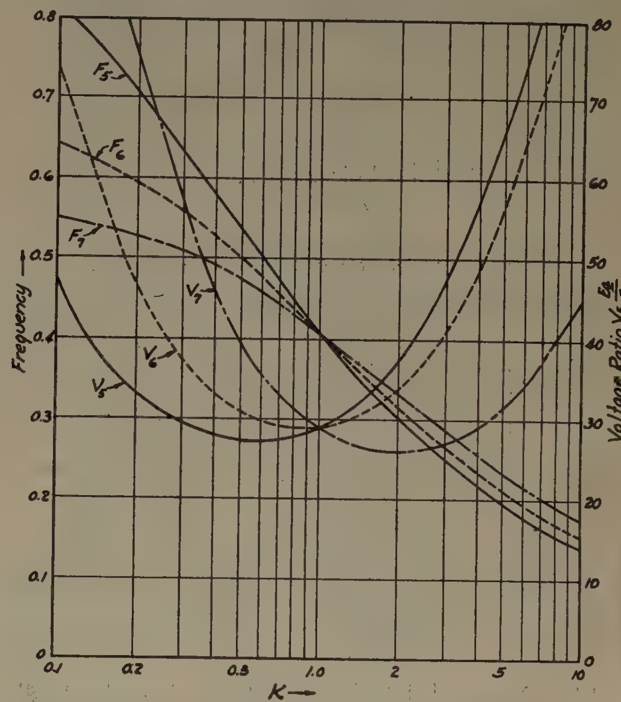


Fig. 5—Frequency and voltage ratio characteristics of three-step ladder network in Fig. 3B.

When oscillating, the shift is 180 degrees and the ($j$) term zero.

$$X^2 = 5K_5 + 1 \quad (7)$$

$$\text{and} \quad F_5 = 1/X = 1/\sqrt{5K_5 + 1}. \quad (8)$$

This is plotted as $F_5$ in Fig. 5. By substituting this value into (6) the plot of $V_5$ is obtained. The curves $V_6$, $F_6$ and $V_7$, $F_7$ are obtained by using either $K_6$ or $K_7$ as the variable in (5).

The curves in Figs. 4 and 5 will give the frequency range, and the required change in amplifier gain to cover this range, when varying any one resistor of either a four-step or a three-step ladder. In all cases the network-voltage ratio passes through a broad minimum value, an operating point where wide frequency shifts may be obtained with only a small amount of amplitude

modulation. These points of minimum loss generally occur at the value of $K$ for which the rate of change of $F$ with $K$ is largest, a fortunate condition for obtaining large frequency changes.

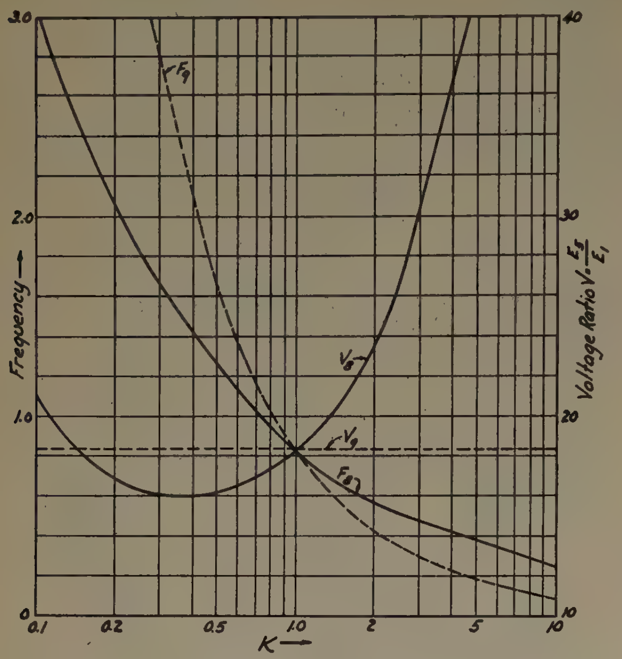


Fig. 6—Effect of varying first and second resistors together, and of all four resistors together, of network in Fig. 3A.

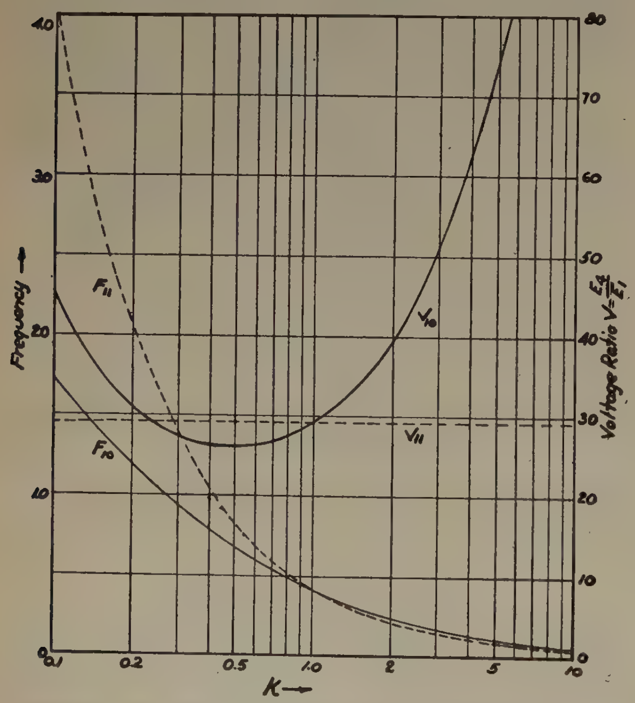


Fig. 7—Effect of varying first and second resistors together, and of all three resistors together, of network in Fig. 3B.

It is possible to control more than one resistor and obtain still wider frequency swings without undue amplitude change. If the first two resistors in the four-step ladder are changed, the curves $F_8$, $V_8$ of Fig. 6 are obtained. When all resistors are varied together the voltage ratio remains constant at 18.36, and the possible swing is theoretically infinite. This is shown by curves $F_9$, $V_9$ in Fig. 6.

When the first two resistors of the three-step ladder are controlled, the curves $F_{10}$, $V_{10}$ of Fig. 7 are obtained; and when all three are controlled, $F_{11}$, $V_{11}$. In this latter case the loss ratio remains constant at 29.

### C. The Zero Shift Network

For the sake of comparison, similar curves for the zero phase-shift network in Fig. 3C are derived and plotted[10] in Fig. 8. Here again $R=1$, and the frequency ordinate must be multiplied by $1/2\pi RC$ to obtain cycles per second. The fundamental equation for $V=E_2/E_1$ is, in this case,

$$V = \frac{X(K_{13} + 2K_{12}) + jK_{12}K_{13} - jX^2}{K_{12}X}. \tag{9}$$

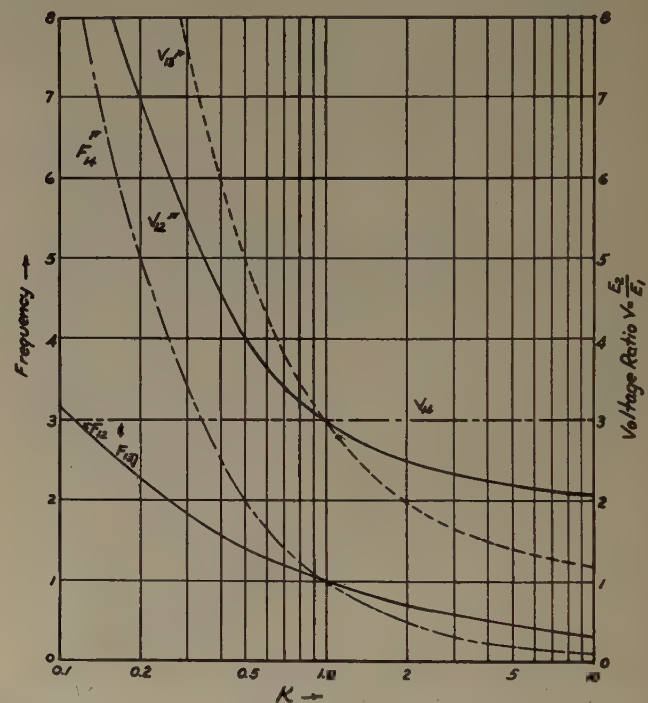


Fig. 8—Frequency and voltage ratio characteristics of network in Fig. 3C when either or both resistors are varied.

For oscillation, the shift is 0 degrees, so the ($j$) terms must cancel:

$$K_{12}K_{13} = X^2. \tag{10}$$

If $K_{12}$ alone is varied and $K_{13}=1$:

$$X^2 = K_{12}$$

$$F_{12} = 1/\sqrt{K_{12}} \tag{11}$$

and the voltage ratio becomes

$$V_{12} = 1 + 2K_{12}/K_{12}. \tag{12}$$

Similarly, $V_{13}$ and $F_{13}$ are obtained as

$$F_{13} = 1/\sqrt{K_{13}} \quad \text{(same as } F_{12}\text{)} \tag{13}$$

and

$$V_{13} = K_{13} + 2/K_{13} \tag{14}$$

and if both resistors are varied together, the equations

[10] See footnote references 4, 5, and 6 for circuits of this type of oscillator.

for $F_{14}$ and $V_{14}$ are obtained:

$$F_{14} = 1/K_{14} \tag{15}$$

$$V_{14} = 3. \tag{16}$$

It can be seen that neither of the amplitude curves $V_{12}$ and $V_{13}$ has a minimum value, but approach, respectively, the limits of 2 and 1. Thus, there is no operating point where a change in frequency may be obtained without at the same time introducing some amplitude change.

Experience has shown that the circuit in Fig. 1 can be operated without noticeable amplitude modulation over ranges where the network loss does not change over 20 per cent. (See topic on Amplitude Characteristics.) The results of the curves in Figs. 4, 5, 6, 7, and 8, therefore, can be tabulated to compare the frequency shift at an amplitude change of 20 per cent. This tabulation also indicates the center operating position, in relative frequency, and the central value of $K$, for determining the ratio of control-tube-circuit resistance to the network resistors.

TABLE I

FREQUENCY SHIFT WHEN 20 PER CENT CHANGE IN NETWORK VOLTAGE RATIO IS ALLOWED

| Resistor Varied | Voltage Ratio | | Mid-Frequency | Percentage Frequency Shift on Mid-Frequency | Mid-Frequency Value of $K$ |
|---|---|---|---|---|---|
| | Minimum | Maximum | | | |
| *A. Four-Step Ladder* | | | | | |
| 1st Res. ($K_1$) | 16.6 | 19.92 | 1.14 | ±38.5 | 0.38 |
| 2nd Res. ($K_2$) | 18.2 | 21.84 | 0.87 | ±31.0 | 0.63 |
| 3rd Res. ($K_3$) | 18.33 | 22.0 | 0.82 | ±24.4 | 0.66 |
| 4th Res. ($K_4$) | 17.0 | 20.4 | 0.67 | ±22.4 | 2.60 |
| 1st & 2nd ($K_8$) | 16.0 | 19.2 | 1.70 | ±56.5 | 0.29 |
| *B. Three-Step Ladder* | | | | | |
| 1st Res. ($K_5$) | 27.0 | 32.4 | 0.53 | ±32.0 | 0.50 |
| 2nd Res. ($K_6$) | 29.0 | 34.8 | 0.43 | ±27.0 | 0.75 |
| 3rd Res. ($K_7$) | 26.0 | 31.2 | 0.333 | ±30.0 | 2.0 |
| 1st & 2nd ($K_{10}$) | 26.0 | 31.2 | 0.75 | ±55.3 | 0.40 |
| *C. Zero Phase-Shift Network* | | | | | |
| *1st Res. ($K_{11}$) | 2.7 | 3.3 | 1.0 | ± 8.0 | 1.0 |
| *1st Res. ($K_{11}$) | 1.2 | 1.47 | 0.398 | ±26.0 | 6.5 |
| *2nd Res. ($K_{12}$) | 2.7 | 3.3 | 1.0 | ±15.0 | 1.0 |
| *2nd Res. ($K_{12}$) | 2.1 | 2.56 | 0.566 | ±44.2 | 3.0 |

* For these curves no minimum value of $V$ exists, so two values are given, one for the center of the range where $K$ is 1, and the other where $K$ is 10 at the low frequency end of the swing.

This tabulation will indicate the amount of frequency swing that can be obtained with a 20 per cent change in amplitude, but it does not indicate how linear these swings will be.

## Amplitude Characteristics and Harmonic Distortion

Narrow frequency shifts can be obtained with very little amplitude change if the center operating positions are chosen to correspond to the minimum $V$ values. When wide limits of shift are desired, some precautions are necessary. There are two obvious methods of reducing this amplitude change, by limiters or by some form of automatic volume control. The second method is applicable where the modulating frequencies are a very small percentage of the carrier frequency. When the modulating frequencies approach the frequency of the carrier, then the automatic control must become a cycle-by-cycle device, or in reality a limiter, if the higher modulating frequencies are not suppressed. In the case of facsimile, where the modulating frequencies are as high as 30 or 40 per cent of the mid-carrier frequency, limiting is the logical method.

One of the attributes of the ladder-type oscillators, as compared to the zero shift type, is its ability to limit without causing high harmonic distortion, or squaring of the resulting output. This limiting is accomplished in the oscillator-tube circuit itself. If the curve $F_1$ is used, and it is desired to swing the full range of ±38.5 per cent, or from 1.54 to 0.74, the tube-driver circuit is adjusted to give the maximum gain of 19.92. The circuit then will oscillate at the frequencies of 1.54 or 0.74 at full amplitude. When the center frequency of 1.14 is reached the gain will be 20 per cent too high, and some squaring of the wave should result. However, the network will have 180 degrees shift only at the fundamental frequency and will thus serve as its own filter. Higher harmonics will pass through with less phase shift and tend to cancel by degeneration. Lower frequencies than the fundamental will be suppressed by attenuation. Thus, as the network is varied to shift frequency its filtering action is also changed to be most effective at the generated frequency. External filters to remove distortion become unnecessary.

This is not true with the zero shift type of a resistance-capacitance oscillator. Here the output squares up very rapidly with a small amount of overdrive, and the frequency falls. This is due to the low order of filtering in the network. In this case, reasonably constant amplitude is necessary to preserve frequency stability and prevent distortion. The high shifts of ±26 and ±44.2 per cent as tabulated for this network probably cannot be realized except by using automatic volume control, or by allowing the oscillator to overdrive at the higher frequencies, and then filtering the harmonics from the output.[11] When overdriven too far, the circuit begins to function as a multivibrator and frequency stability is lost.

## Linearity

In the tube circuit furnishing the variable resistance, considering the tube $T_4$ as a perfect cathode follower, the resistance inserted will be half that of either tube. The usual method of expressing the tube characteristics in a formula is

$$i_p = (E_p + \mu E_g)/R_{p_0}. \tag{17}$$

The variable resistance inserted, therefore, will be

$$R_v = E_p R_{p_0}/2(E_p + \mu Eg). \tag{18}$$

In this case, due to the action of $T_4$, $E_p$ is always the same and equal to one half $E_B$.

Therefore, $$R_v = E_B R_{p_0}/2(E_B + 2\mu Eg). \tag{19}$$

When all of the resistors of a network are varied at the same time, the frequency is proportional to $1/K$.

[11] It should be noted that in some facsimile uses the range is over two to one from maximum to minimum frequency. Thus, any external filtering of the second harmonic of the lowest frequency would also eliminate the upper-frequency fundamental, and introduce serious amplitude distortion.

Therefore, if $R_v$ is made equal to $K$ by the proportionality constant $T$

$$F_4 = T \cdot 2(E_B + 2\mu Eg)/E_B R_{p_0}. \quad (20)$$

Thus, frequency is a linear function with the control voltage on the grid of $T_3$.

When one or two resistors only are varied, the relationship does not hold, but the $K \cdot F$ curve of the network can usually be made to follow closely the $R_p \cdot Eg$

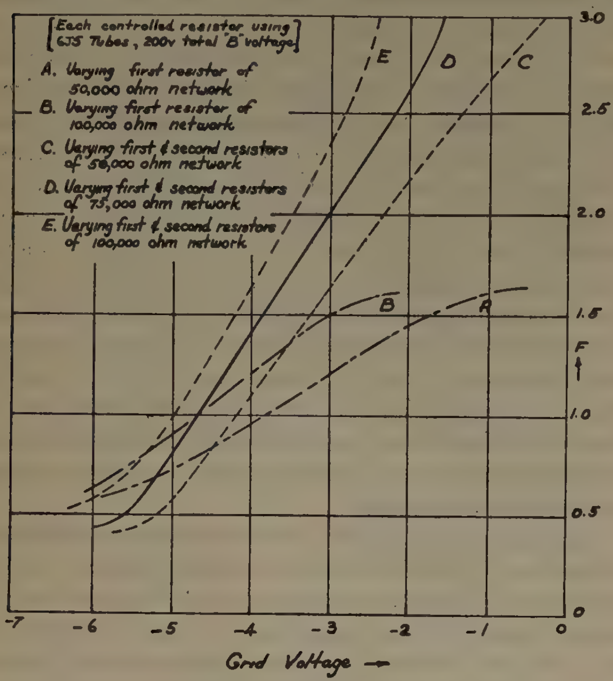


Fig. 9—Calculated characteristics using 6J5 control tubes to vary one or two resistors of network in Fig. 3A.

curve of the tube over wide ranges. Referring to the experimental curves in Fig. 2, it can be seen that the linear portion of $B$ occupies about 80 per cent of the total swing. In certain experimental modulated oscillators this percentage was increased to over 90 per cent by using a pentode for $T_3$.

Fig. 9 illustrates how this fitting of tube curve and network may be carried out to obtain a frequency swing linear with grid voltage. For curve $A$, the network resistance in each step is assumed to be 50,000 ohms, and the two control tubes $T_3$ and $T_4$ are 6J5's. The total B voltage is assumed as 200 volts. If only the first resistor is variable, then the curve $F_1$, Fig. 4, is the network characteristic, and the points on curve $A$ are determined as follows: At the point where $F$ is 1, from curve $F_1$, Fig. 4, the value of $K$ for the variable resistor is found to be 0.57. The network resistors that are not varied were assumed 50,000 ohms, so the two tubes (together) of the control must show $K$ (50,000) or 28,500 ohms. As the two tubes are effectively in parallel, each tube must have twice this or 57,000 ohms actual resistance. With a total B supply of 200 volts, each tube then draws $200/2 \times 57{,}000$ or 1.755 milliamperes. The grid voltage necessary for this plate current with 100 volts on on the plate is read from tube charts or actual test values, $-3.8$ volts in this case. This process is repeated for as many points as desired. The curves where two resistors are varied are obtained in the same manner but using the $F_8$ curve in Fig. 6 to represent the network. Both $A$ and $B$ curves are S shaped, but $B$ has a considerably more nearly straight center section. This indicates the 6J5-tube curve is better suited in the low-current region.

The curves $C$, $D$, and $E$ show how the correct network resistance can be chosen for greatest linearity when two resistors are varied. Curve $C$ has a long S shape, but is much more nearly straight than was found for only one variable resistor as in $A$. $D$ is almost perfect over the range of from $F=0.5$ to $F=2.5$, a shift of $\pm 67$ per cent about a center frequency of 1.5. Further increase in the network resistors gives line $E$, which has shifted to a hyperbolic shape and is curved throughout its length.

In using the curve $D$, these limits of linearity, $F=0.5$ to $F=2.5$, are put back into the voltage ratio curve $V_8$, Fig. 6, to see if this swing can be obtained without amplitude modulation. It is found that the $F=2.5$ end can be reached with a gain increase of only 14.4 per cent over the minimum value, but the $F=0.5$ end of the sweep requires a gain increase of 87 per cent. Obviously, this lowest frequency cannot be reached without more amplitude modulation than can be corrected by the natural limiting. The low-frequency end of the swing should be restricted to about $F=0.8$ for a symmetrical mid-carrier position. In this manner a linear swing of $\pm 51.5$ per cent can be realized with only 14.4 per cent increase in gain, a value easily within the range of the limiting and, therefore, not causing distortion.

## Conclusions

Direct-frequency modulation of resistance-capacitance oscillators is shown to be a very practical means of obtaining large swings on relatively low frequency carriers. The method is especially adaptable for facsimile and similar communication systems.

The circuits are simpler than the beat oscillators formerly used, and are much more stable in adjustment. Spurious amplitude modulation can be reduced to a negligible amount by choice of the proper network and constants, and the harmonic content of the output signal is low. Modulation of the resistance-capacitance oscillator is accomplished by a pair of control tubes so balanced that no transients or components of the original signal appear in the output.

Facsimile systems have been put in operation with these circuits with frequency swings of 500 to 1500 cycles, 2000 to 4000 cycles, 4000 to 9000 cycles, and 16,000 to 24,000 cycles.

Design of such systems has been simplified by providing the $F$ versus $K$ curves. These facilitate the choice of network constants and control tubes for any desired frequency swing.

# Current Stabilizers*

J. N. VAN SCOYOC†, ASSOCIATE, I.R.E., AND E. H. SCHULZ†, ASSOCIATE, I.R.E.

***Summary*—The degree of regulation to be expected from several types of current stabilizers is considered. Regulation equations are derived and a graphical-design method is presented.**

**The regulation produced by a single pentode is much better than that of a single triode. Two cascode-connected pentodes give a better regulation but the tube drop is greater and the circuit is more complex. The regulation may be improved by using nonlinear resistors in the control circuit.**

## Introduction

Current stabilizers are used in various applications to regulate against current changes produced by line-voltage fluctuation or load-resistance variation. These circuits are used in photometry, measurement work, magnetron oscillators, etc. A current stabilizer may be used advantageously as a coupling device in direct-current amplifier work where a voltage divider is required to satisfy quiescent conditions, but no division of voltage change is desired.

It is the purpose of this paper to show the degree of regulation to be expected from several circuits and to present a design procedure. Several practical regulator circuits are described.

## Derivations

The circuit diagram and equivalent circuit of a simple degenerative current stabilizer are shown in Fig. 1. A solution of the equivalent circuit for a change in the

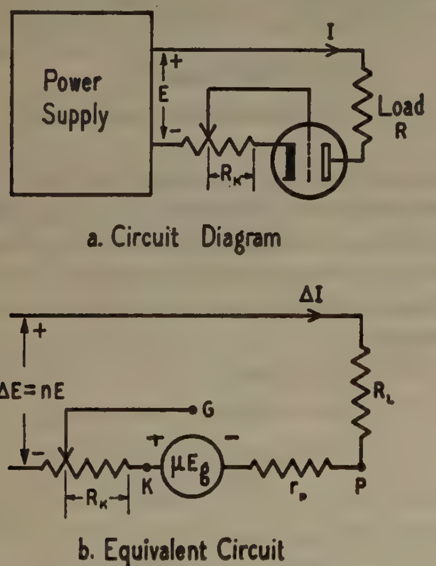


Fig. 1—Cascode-connected current stabilizer.

power-supply voltage yields the following expression for the change in load current:

$$\Delta I = \frac{nE}{R_L + r_P + (1+\mu)R_K} \quad \text{(for } E) \tag{1}$$

where $\Delta I$ is the current change in amperes

$E$ is the power-supply voltage

$n = \Delta E/E$ is the per unit change of voltage.

The current will be

$$I = E_L/R_L \tag{2}$$

where $E_L$ is the voltage drop across the load. The per unit change of current is then

$$\frac{\Delta I}{I} = \left(\frac{R_L}{R_L + r_P + (1+\mu)R_K}\right)\left(\frac{E}{E_L}\right)n \quad \text{(for } \Delta E). \tag{3}$$

The derivation of the regulation expression for changes in load resistance is simplified by assuming the tube-characteristic curves to be straight lines. Then,

$$I = \frac{E}{R_L + r_P + (1+\mu)R_K}. \tag{4}$$

Taking the derivative with respect to $R_L$ and expressing the results in terms of current and resistance increments,

$$\Delta I = \frac{-mR_LE}{[R_L + r_P + (1+\mu)R_K]^2} \quad \text{(for } \Delta R_L) \tag{5}$$

where $m$ is the per unit change in $R_L$ (i.e., $\Delta R_L = mR_L$). If (5) is divided by (2) and the results are simplified, the per unit change in current may be expressed as

$$\frac{\Delta I}{I} = \left(\frac{R_L}{R_L + r_P + (1+\mu)R_K}\right)^2\left(\frac{E}{E_L}\right)m \quad \text{(for } R_L). \tag{6}$$

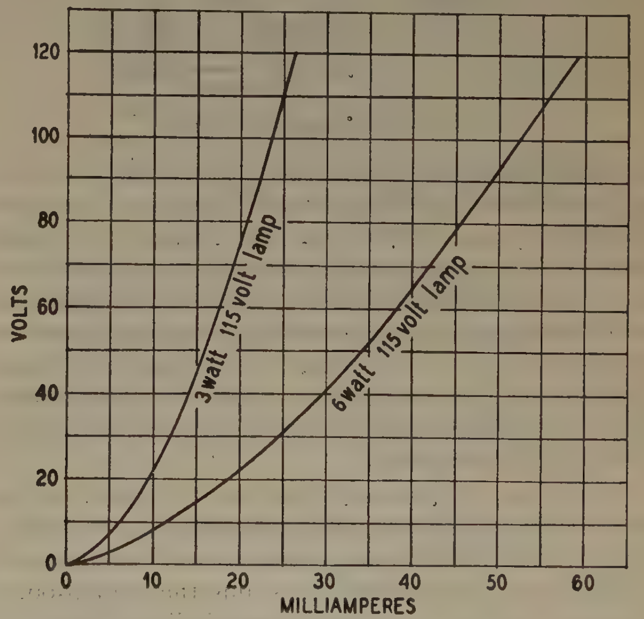


Fig. 2—Volt-ampere characteristics of tungsten lamps suitable for use in current stabilizer.

## Use of Nonlinear Element for $R_K$

An examination of (3) and (6) indicates that $R_K$ should be as large as possible for good regulation. However, increasing $R_K$ increases the grid bias and hence the tube drop. As a result the ratio $E/E_L$ is increased,

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thus limiting the advantage which may be secured in this way. Also the plate dissipation of the tube is increased by the increase in voltage, and the required power-supply voltage is increased.

The use of a nonlinear resistor (such as a tungsten-filament lamp) for $R_K$ will result in some improvement because the tube drop is dependent on the actual value of the cathode resistor, while, as far as changes in current are concerned, $R_K$ is equal to the slope of the volt-ampere curve of the resistor. The slope of this curve may be from two to three times as large as the actual resistance at the operating point (see Fig. 2) and hence an appreciable improvement in regulation may be obtained.

Because lamps and similar thermal devices have a definite time lag the improvement in regulation does not apply to sudden surges or residual hum voltages. The advantage to be gained from their use is, therefore, somewhat limited.

## Pentode Stabilizer Circuits

Equations (3) and (6) indicate that for the best regulation $r_P$ and $\mu$ should be large. Both requirements are

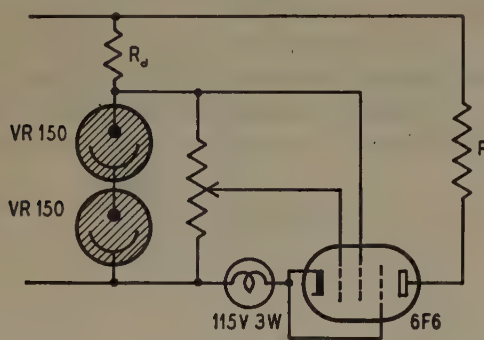


Fig. 3—Pentode current stabilizer.

satisfied by the use of a pentode. Fig. 3 shows the circuit of a stabilizer using a pentode. The screen voltage is supplied by means of $VR$ gas tubes connected in series with a dropping resistor $R_d$ across the line. A portion of the voltage existing across the $VR$ tubes is applied to the control grid so as to reduce the tube drop. However, the change in grid voltage due to a change in $I$ is equal to the change in voltage drop across $R_K$. This scheme makes it possible to use a large value of $R_K$ without an unduly high tube drop.

An increase in load current reduces the screen voltage because the drop across $R_K$ is included in the screen-cathode voltage. This decrease in screen voltage tends to reduce the current and thus aids the regulation. A treatment similar to that used above indicates that

$$\frac{\Delta I}{I}=\left(\frac{R_L}{R_L+r_P+(1+\mu_1+\mu_2)R_K}\right)\left(\frac{E}{E_L}\right)n \text{ (for } \Delta E) \quad (7)$$

$$\frac{\Delta I}{I}=\left(\frac{R_L}{R_L+r_P+(1+\mu_1+\mu_2)R_K}\right)^2\left(\frac{E}{E_L}\right)m \text{ (for } \Delta R_L) \quad (8)$$

where $\mu_1$ is the control-grid—plate-amplification factor and $\mu_2$ is the screen—plate-amplification factor. This derivation neglects the effects of screen current through $R_K$ and of changes in screen current which will actually reduce the regulation somewhat; however, if the screen current is not too large, the results will be reasonably accurate.

## Cascode-Connected Stabilizer

Fig. 4 shows a regulator with two tubes connected in cascode[1] and its equivalent circuit. The various grid

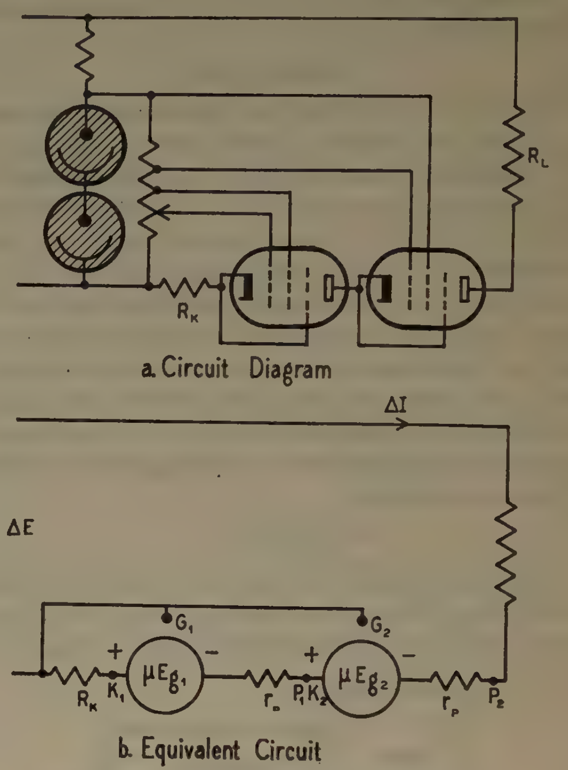


Fig. 4—Cascode-connected stabilizer.

potentials are obtained from a voltage divider connected across two (or more) $VR$ tubes in series. Since each of the grid voltages consists of a constant voltage plus the drop across $R_K$, the change in each grid voltage is equal to the change in drop across $R_K$. The following equations may be written:

$$\begin{aligned} E_{g1} &= -\Delta I R_K \\ E_{g2} &= -\Delta I R_K - \Delta I r_P + \mu E_{g1} \\ \Delta I(R_L + R_K + 2r_P) &= \Delta E \\ I &= E_L/R_L. \end{aligned} \quad (9)$$

The solution of these equations yields

$$\Delta I=\left(\frac{R_L}{R_L+(2+\mu)r_P+(1+\mu)^2R_K}\right)\left(\frac{E}{E_L}\right)n \text{ (for } \Delta E) \quad (10)$$

A process similar to that used in the derivation of equation (6) gives

$$\Delta I=\left(\frac{R_L}{R_L+(2+\mu)r_P+(1+\mu)^2R_K}\right)^2\left(\frac{E}{E_L}\right)m \text{ (for } \Delta R_L) \quad (11)$$

[1] The term "cascode" as distinguished from "cascade" applies to an amplifier in which the tubes are connected in series to obtain direct coupling.

The effect of change in screen voltage was ignored in this derivation, but it may be taken into account by substituting $\mu_1+\mu_2$ for $\mu$. However, the effect of change in screen current will tend to reduce the effect of screen-voltage changes and hence the results of (10) and (11) are well within design accuracy.

## Current-Stabilizer Design

The design procedure consists of the choice of a tube, a value of $R_K$, and the power-supply voltage $E$ required to give a certain maximum current variation for given ranges of line voltage and of load resistance.

For small changes in voltage and load resistance the equations given above may be used to calculate the degree of regulation to be expected; however, for large changes it is necessary to use a graphical solution.

Fig. 5 shows a graphical solution of a current stabilizer. For a given value of $R_K$ the grid voltage for any assumed plate current is known and hence a plate-current—plate-voltage curve may be constructed for this value of $R_K$ as follows: Assume a value of $i_b$, calculate the grid voltage, and plot the intersection of the tube curve for this value of grid voltage and the ordinate for the assumed current. Repeat the process for various assumed currents until sufficient points are obtained to plot a curve. If a lamp is used for $R_K$, the grid voltage must be taken from the lamp volt-ampere curve.

After the curve for the assumed $R_K$ is obtained, the current to be expected for any $E$ and $R_L$ may be found by drawing a load line of slope $-\tan^{-1}[1/(R_L+R_K)]$ through the point $e_b=E$, $i_b=0$. The intersection of this load line and the characteristic curve gives the current to be expected. In Fig. 5a a solution is obtained for a constant load resistance and variable impressed voltage by sliding the load line along the axis over the range of voltages to be expected and reading the values of current. In Fig. 5b a solution is obtained for a constant voltage and variable load resistance by drawing the various load lines through the point $i_b=0$, $e_b=E$ and noting the values of current to be obtained.

It should be noted that this solution ignores the effect of screen-voltage changes and hence gives a pessimistic answer. It may be necessary to plot special curves applying to the particular value of screen voltage to be used.

A tube with high values of $\mu$ and $r_P$ should be chosen. Also the tube must have sufficient current capacity and its plate dissipation capacity must be greater than the maximum plate dissipation to be expected.[2] A sufficient number of $VR$ tubes should be used to obtain a reasonable screen voltage. If it is necessary to operate with a low screen voltage (to economize on $VR$ tubes or because $E$ is not sufficient to keep the glow current within the normal range for variations of $E$) it may be desirable to operate the control grid with a positive potential to reduce the tube drop. The resistor $R^d$ should be of such a size as to keep the $VR$ tube current within the normal range.

[2] Maximum plate dissipation is obtained when $E$ is a maximum and $R_L$ a minimum. The plate dissipation is $E_bI$ where $E_b$ is the distance $oa$ in Fig. 5b.

## Comparison of Stabilizers and Conclusions

Table I indicates the degree of regulation to be expected from different circuits supplying the same load. The values in this table were calculated and are accurate

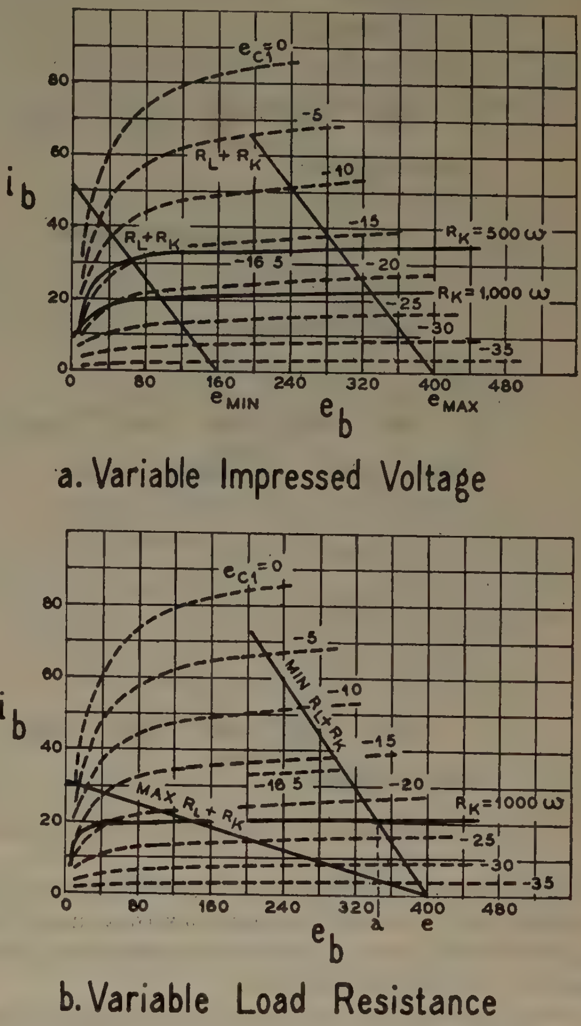


Fig. 5—Graphical solution for a current stabilizer.

for small changes only. However, in most cases the changes to be regulated against are less than 10 to 20 per cent and these results are reasonably accurate. Fig.

TABLE I
CALCULATED CHARACTERISTICS OF SEVERAL CURRENT STABILIZERS USED TO SUPPLY 15 MILLIAMPERES TO A 3000-OHM LOAD

| Circuit | $\Delta I/I$ for $\Delta E=nE$ | $\Delta I/I$ for $\Delta R_L=mR_L$ |
|---|---|---|
| 1. 6F6 triode (Fig. 1) $R_K$=3-watt lamp, $E$=550 volts | $0.9\ n$ | $-0.06\ m$ |
| 2. 6F6 triode (Fig. 1) $R_K$=6-watt lamp, $E$=245 volts | $0.93\ n$ | $-0.16\ m$ |
| 3. 6F6 pentode (Fig. 3) $R_K$=3-watt lamp, $E$=270 volts $E_c=-25$ volts, $E_{sg}=250$ volts (2-$VR$150 tubes) | $0.023\ n$ | $-6\times10^{-5}\ m$ |
| 4. 6F6 cascode (Fig. 4) $R_K$=3-watt lamp, $E$=550 volts $E_c=-17.5$ volts, $E_{sg}=220$ volts (3-$VR$150 tubes) | $1.87\times10^{-4}\ n$ | $2.86\times10^{-9}\ m$ |

6 shows experimetal curves taken in the laboratory on regulators similar to those in Table I with the exception that the screens were operated at a lower voltage and the control grids were operated with a small negative or

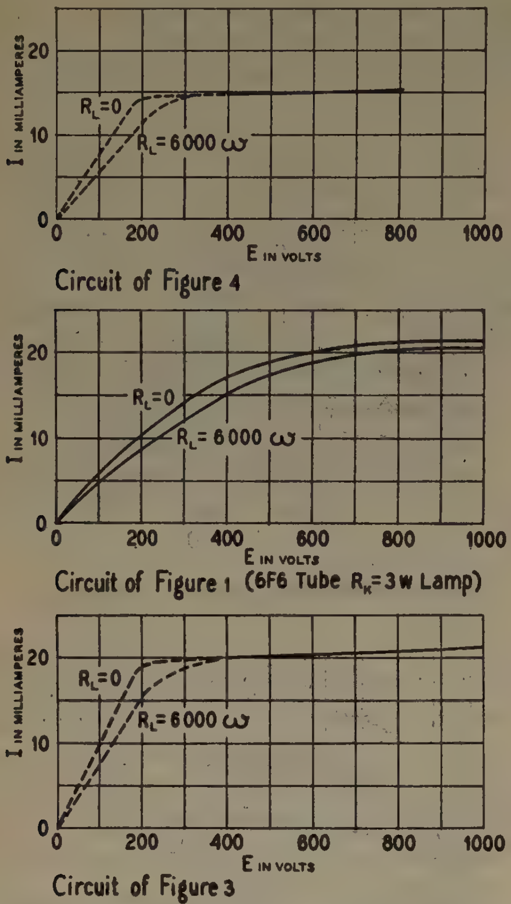


Fig. 6—Experimental results on three current-stabilizer circuits.

perhaps a positive grid voltage. It is evident from these results as well as from the equations that much better regulation may be obtained for load-resistance variation than for line-voltage variation.

Many applications of these circuits suggest themselves. For example, the use of a current stabilizer as a dropping resistor in series with a *VR* tube-voltage regulator will make it possible to maintain the *VR* tube current within the normal range over a much wider range of voltage variation.

Fig. 7 shows a current stabilizer used as an element in the voltage divider feeding the grid of the direct-current amplifier in a degenerative voltage stabilizer. This divider is necessary to keep the grid at a negative potential but a resistance divider is undesirable because it also divides the changes of output voltage. Since the current in the divider remains constant and hence the drop across the resistor is held constant, any change in out-

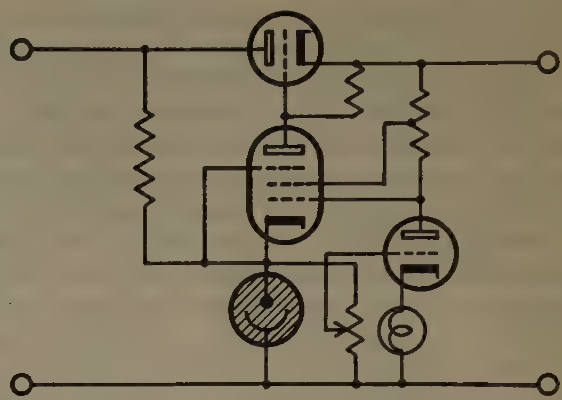

Fig. 7—Current stabilizer used as voltage divider in a degenerative-type voltage regulator.

put voltage must appear across the regulator tube and hence upon the grid of the direct-current amplifier.

It has been shown that the stabilizers in the order of their regulating capability are: (1) cascode-connected circuit, (2) pentode circuit, and (3) triode circuit. If the stabilizers were listed in the order of their relative simplicity, the order would be reversed. The triode circuit gives the poorest regulation and has the largest tube drop. The cascode connection gives the best regulation but the tube drop is high. The pentode is a good compromise between good regulation and simplicity and it gives the smallest tube drop.

### Reference

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# Noise Figures of Radio Receivers*

H. T. FRIIS†, FELLOW, I.R.E.

**Summary—A rigorous definition of the noise figure of radio receivers is given in this paper. The definition is not limited to high-gain receivers, but can be applied to four-terminal networks in general. An analysis is made of the relationship between the noise figure of the receiver as a whole and the noise figures of its components. Mismatch relations between the components of the receiver and methods of measurements of noise figures are discussed briefly.**

## Introduction

THE importance of noise originating in a radio receiver has increased as shorter and shorter wavelengths have come into practical usage. Many papers on the subject, notably those by Llewellyn[1] and Jansky,[2] have appeared since the writer, in 1928, showed experimentally[3] that thermal-agitation noise (Johnson noise) determined the absolute sensitivity of short-wave radio receivers. Early in 1942 North[4] suggested the adoption of a standard for the absolute sensitivity of radio receivers which differed by a factor of 2 from the standard used by us at that time. We adopted his standard, since ours was somewhat limited in that it was based on matched impedances in the input circuit of the receiver.

In this paper a more rigorous definition of the standard of absolute sensitivity, the so-called noise figure, of a radio receiver is suggested. The definition is not limited to high-gain receivers, but can be applied to four-terminal networks in general. It also makes it possible by a simple analysis to give the relationship between the noise figure of the receiver as a whole and the noise figures of its components. In the case of a double-detection receiver these components may be a high-frequency amplifier, a frequency converter, and an intermediate-frequency amplifier. The paper also gives a brief description of methods of measurements of noise figures.

The four-terminal network whose noise figure is to be defined is shown schematically in Fig. 1. A signal generator is connected to its input terminals and an output circuit is also indicated. The input and output impedances of the network may have reactive components and they may be matched or mismatched to the generator and the output circuit, respectively. The four-terminal network may be, for instance, an amplifier, a converter, an attenuator, or a simple transformer. The presence of the signal generator is required for the definition that follows, but the attenuator in the signal generator and the output circuit toward the right are shown only to illustrate measurements of noise figures and gains.

The noise figure will be defined in terms of available signal power, available noise power, gain, and effective bandwidth. The definitions of these terms will be given and discussed next.

## Available Signal Power

A generator with an internal impedance $R_0$ ohms and electromotive force $E$ volts delivers $E^2R_1/(R_0+R_1)^2$ watts into a resistance $R_1$ ohms. This power is maximum and equal to $E^2/4R_0$ when the output circuit is matched to the generator impedance, that is when $R_1=R_0$. $E^2/4R_0$ is hereafter called the available power of the generator, and it is, by definition, independent of the impedance of the circuit to which it is connected. The output power is smaller than the available power when $R_1$ is unlike $R_0$, since there is a mismatch loss. In amplifier input circuits a mismatch condition may be beneficial[5] due to the fact that it may decrease the output noise more than the output signal. It is the presence of such mismatch conditions in amplifier input circuits that makes it desirable to use the term available power in this paper. The symbol $S_g$ will be used for the available signal power at the output terminals of the signal generator shown in Fig. 1. $S_g$ is here equal to $V^2/RA$ watts where $V$ is the voltage across the input terminals

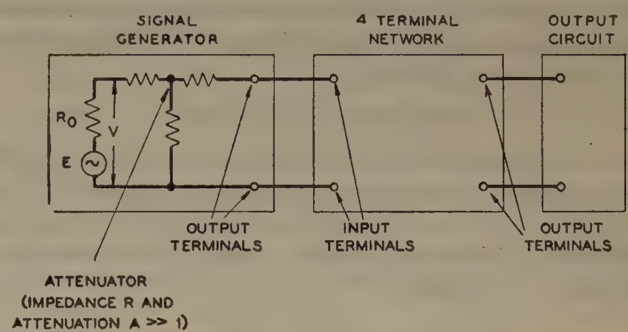


Fig. 1

of the attenuator, $R$ the characteristic impedance of the attenuator, and $A$ the nominal attenuation ($A$ is assumed large).

The output terminals of any network may be considered as being the output terminals of a signal generator. The symbol $S$ will be used for the available signal power at the output terminals of the four-terminal network shown in Fig. 1.

* Decimal classification: R261.2. Original manuscript received by the Institute, September 7, 1943.

† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] F. B. Llewellyn, "A study of noise in vacuum tubes and attached circuits," Proc. I.R.E., vol. 18, pp. 243–266; February, 1930.

[2] K. G. Jansky, "Minimum noise levels obtained on short-wave radio receiving systems," Proc. I.R.E., vol. 25, pp. 1517–1531; December, 1937.

[3] Unpublished report.

[4] D. O. North, "The absolute sensitivity of radio receivers," *RCA Rev.*, vol. 6, pp. 332–344; January, 1942. The reader is referred to this paper for reference to other papers on this subject.

[5] That such an improvement might be possible was first discussed in detail by F. B. Llewellyn in his paper "A rapid method of estimating the signal-to-noise ratio of a high gain receiver." Proc. I.R.E., vol. 19, pp. 416–421; March, 1931.

### Gain

The gain of the network is defined as the ratio of the available signal power at the output terminals of the network to the available signal power at the output terminals of the signal generator. Hence

$$G = S/S_g. \tag{1}$$

This is an unusual definition of gain since the gain of an amplifier is generally defined as the ratio of its output and input powers. This new definition is introduced here for the same reason that made it desirable to use the term available power. Note that while the gain is independent of the impedance which the output circuit presents to the network, it does depend on the impedance of the signal generator.

The four-terminal network has generally some kind of band-pass characteristic. The gain $G$ is defined as that at the mid-band frequency.

### Available Noise Power

As in the case of signal power, the available noise power between two terminals is defined as the noise power which would be absorbed by a matched output circuit.

The symbol $N$ will be used for the available noise power at the output terminals of the network. This power is due to all the noise sources in the network itself and the Johnson-noise sources in the signal generator, but noise sources in the output circuit toward the right in Fig. 1 are not included.

The Johnson-noise power available from a resistance will be discussed now. Any resistance, $R$ ohms, acts as a Johnson-noise generator with a mean-square electromotive force equal to $4KTRdf$. $K$ is Boltzmann's constant $=1.38\times10^{-23}$, $T$ is the absolute temperature of the resistance, and $df$ is the bandwidth. The available Johnson-noise power is then

$$4KTRdf/4R = KTdf \text{ watts} \tag{2}$$

and this is the noise power available over the band $df$ at the output terminals of the signal generator in Fig. 1. It is, in fact, the available noise power between any two terminals of a passive network when all its parts have the same temperature $T$.

### Effective Bandwidth

The contribution to the available output noise by the Johnson-noise sources in the signal generator is readily calculated for an ideal or square-top band-pass characteristic and it is $GKTB$ where $B$ is the bandwidth in cycles per second. In practice, however, the band is not flat; i.e., the gain over the band is not constant but varies with the frequency. In this case, the total contribution is $\int G_f KTdf$ where $G_f$ is the gain at the frequency $f$. The effective bandwidth $B$ of the network is defined as the bandwidth of an ideal band-pass network with gain $G$ that gives this contribution to the noise output. Therefore,

$$GKTB = \int G_f KTdf$$

or

$$B = \frac{1}{G}\int G_f df. \tag{3}$$

### Noise Figures

The noise figure of the network in Fig. 1 will now be defined in terms of $S_g$, $S$, $KTB$, and $N$.

It is important to have the highest possible signal-to-noise ratio at the output terminals of the network. The maximum value of this ratio would be as high as the available-signal-to-available-noise ratio at the signal-generator terminals if there were absolutely no noise sources present in the network. Simple lossless transformers or filters are examples of networks with no noise sources. In general, however, noise sources are present and these noise sources reduce the available signal-to-noise ratio at the output terminals of the network. The noise figure $F$ of the network[6] is defined as the ratio of the available signal-to-noise ratio at the signal-generator terminals to the available signal-to-noise ratio at its output terminals.[7] Thus

$$F = (S_g/KTB)/(S/N) = (S_g/KTB)(N/S) \tag{4}$$

and since

$$G = S/S_g$$
$$F = (1/G)(N/KTB). \tag{5}$$

Solving for $N$ gives the following expression for the available noise output:

$$N = FGKTB \text{ watts}. \tag{6}$$

This noise output includes the contribution made by the Johnson-noise source in the signal generator. This contribution is $GKTB$. The available output noise due only to noise sources in the network is, therefore,

$$(F-1)GKTB \text{ watts}. \tag{7}$$

All the terms in (4), (5), (6), and (7) have been defined, but a value for the temperature $T$ of the generator terminal impedance must still be chosen before the noise figure is definite. It is suggested that the noise figure be defined for a temperature of 290 degrees Kelvin (63 degrees Fahrenheit). Then

$$KT = 1.38\times10^{-23}\times290$$
$$= 4\times10^{-21} \text{ watts per cycle bandwidth}.$$

The relationship between the noise figure and the degree of mismatch that exists between the network and its output and input circuits is important. Definition (4) shows clearly that the output circuit and hence its coupling with the network has no effect on the value of the noise figure. However, it also shows that the noise figure does depend on the degree of mismatch between the generator and the network since both $S$ and $N$ will vary with the magnitude of this mismatch.

[6] We have, until now, used the symbol $\overline{NF}$ for the noise figure, but we shall use, hereafter, the symbol $F$ suggested by Dr. S. Roberts.

[7] Because of this definition, the noise figure has also been called "excess noise ratio."

### Measurement of the Noise Figure

Although a detailed description of noise-figure-measuring equipment will not be given in this paper, it is believed worth while to outline a method of such measurements.

It is not difficult to measure the noise figure $F$ when the gain of the network is so large that a noise-power reading can be obtained by means of a thermocouple connected between the output terminals of the network (Fig. 1). The measurement procedure is then simply to adjust the attenuation $A$ of the signal attenuator until the output reading is double that due to noise only which is obtained with the signal generator turned off. $S$ is then equal to $N$ and definition (4) gives

$$F = S_g/KTB = V^2/4RAB \times 10^{-21}. \qquad (8)$$

The effective band $B$ is calculated from a gain-versus-frequency curve. The voltage $V$ across the input terminals of the signal attenuator may be measured by means of thermocouples, tube voltmeters, thermistors, etc., and by cross-checking such different equipment the value of $V$ can be obtained with adequate accuracy even in the centimeter-wavelength range. It is more difficult to obtain an accurate value of the attenuation $A$ because of its large magnitude. For a short-wave receiver for which $F$ may be as small as 3, formula (8) gives $A = 5.2 \times 10^{13}$ for $R = 80$ ohms, $V = 1$ volt, and $B = 20{,}000$ cycles. Only very careful work will give satisfactory data with such magnitudes of attenuation. Very thorough shielding of the signal generator is one important requirement.

The noise figure of a network made up of nondissipative elements is unity since it contains no noise sources (expression (7) is equal to zero). The losses in simple transformers and filters are generally sufficiently low to come under this classification. The noise figure of an attenuator at 63 degrees Fahrenheit temperature is by (5) equal to its attenuation when it is matched to the signal generator since under these conditions $N = KTB$ and attenuation $= 1/G$. A network made up of a transmitting and a receiving antenna is equivalent to an attenuator with an attenuation $A$ equal to the ratio of transmitted to received power. Assuming no static or star noise and no circuit losses in the receiving antenna, its noise figure is, by (5), $F = A(N/KTB)$. If $T_r$ is the absolute temperature of the radiation resistance of the receiving antenna, $N = KT_rB$. Hence $F = A(T_r/T) = A(T_r/290)$. The value of $T_r$ is not definitely known, but $T_r = T$ is believed to be a good approximation for antennas whose radiation strikes the earth.[8]

### Noise Figures for Two Networks in Cascade

If the gain of the network shown in Fig. 1 is not large, an amplifier following the network is required to obtain a noise-output reading. For this case a noise-figure analysis of two networks in cascade is required. Also from a design point of view it is important to know the relationship between the noise figure of a whole receiver and the noise figures of its components since it indicates the component on which efforts for improvement are worth while.

The two networks are shown schematically in Fig. 2. We are also considering here the general case where the two networks, the generator, and the output circuit may be either matched or mismatched. The definitions given for a single network will now be applied to the network $ab$ made up of the two networks $a$ and $b$ in cascade and to the individual networks $a$ and $b$.

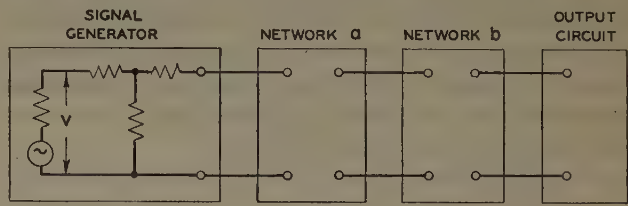


Fig. 2

Expression (6) gives for the available output noise at the output terminals of network $b$

$$N_{ab} = F_{ab}G_{ab}KTB_{ab} \text{ watts.} \qquad (9)$$

To simplify the analysis, it will be assumed that the two networks have the same ideal or square-top band-pass characteristics ($B_a = B_b = B$). The equivalent band $B_{ab}$ is then equal to $B$. The total gain $G_{ab}$ is by the definition of gain (1) equal to $G_aG_b$. Then

$$N_{ab} = F_{ab}G_aG_bKTB \text{ watts.} \qquad (10)$$

A new expression for this noise power may be derived by applying expression (6) to network $a$ and expression (7) to network $b$. Applying (6) to network $a$, the available noise power at its output terminals is

$$N_a = F_aG_aKTB \text{ watts.} \qquad (11)$$

Multiplying this power by $G_b$ gives then the following expression for the available noise power at the output terminals of network $b$ due to the noise sources in network $a$ and the Johnson-noise sources in the signal generator

$$F_aG_aG_bKTB \text{ watts.} \qquad (12)$$

Expression (7) applied to network $b$ gives the following expression for the available noise power at the output terminals of network $b$ due to noise sources in network $b$ only,

$$(F_b - 1)G_bKTB \text{ watts.} \qquad (13)$$

The total available noise power $N_{ab}$ at the output terminals of network $b$ is the sum of the noise powers given by (12) and (13), hence,

$$\begin{aligned} N_{ab} &= F_aG_aG_bKTB + (F_b - 1)G_bKTB \\ &= \left(F_a + \frac{F_b - 1}{G_a}\right)G_aG_bKTB. \end{aligned} \qquad (14)$$

Comparing this expression with (10) gives the following simple relationship between the noise figures of the two networks

$$F_{ab} = (F_a + F_b) - 1/G_a. \qquad (15)$$

[8] For further information on this subject the reader is referred to a paper by R. E. Burgess, "Noise in receiving aerial systems," *Proc. Phys. Soc.*, vol. 53, pp. 293–304; May, 1941.

This relationship is valid for any distribution of noise power throughout the bands of the two networks. The assumptions made in regard to the band-pass characteristics could be made less severe for uniform noise-power distribution. Although they do not seriously limit the usefulness of relationship (15) for practical cases, it is recommended that both the effect of nonuniform noise distribution and the effect of unequal and nonideal bands be studied to make sure whether it is necessary to use correction factors for the different terms of (15).

The rather complicated matter of the effects of nonuniform noise distribution and nonideal band characteristics may be clarified somewhat by pointing out that the relationship (15) may always be applied to an element of band $df$ at a frequency $f$ in the band of the networks. It usually will be found that the noise figure corresponding to an element of band varies somewhat across the band of an actual network.

The noise figures $F_a$ and $F_b$ in (15) will be discussed next. Network $b$ has no effect on the noise figure $F_a$ of network $a$. This follows from the discussion of a single network. That discussion also pointed out that network $a$ does affect the noise figure $F_b$ of network $b$. Therefore, if $F_b$ is measured separately by a signal generator, as shown in Fig. 1, then this signal generator must have a terminal impedance which is identical to the output impedance between the output terminals of network $a$.

### Mismatch Relations for Two Networks

The reader is referred to a paper by Burgess[8] and a more recent paper by Herold[9] for detailed discussions of the advantage of mismatch relations, and only a brief discussion will be given here.

The over-all noise figure for two networks has a minimum value when the degree of mismatch between them is made identical to the mismatch which gives the lowest noise figure for the second network when it is connected directly to a signal generator. Offhand, this is not evident, but an analysis of the matching condition between a signal generator and a network shows that the optimum matching condition is independent of any noise sources in the signal generator. For the lowest over-all noise figure, the optimum matching condition between the signal generator and network $a$ does, on the other hand, depend on both networks. When network $a$ is a low-gain converter and network $b$ an intermediate-frequency amplifier, the highest possible gain $G_a$ in the first network, which is obtained when it is matched to the signal generator, will in general give the lowest over-all noise figure.

### Noise Figures for Several Networks in Cascade

The analysis for two networks may be easily extended to more than two networks. For example, if three networks are considered, (15) gives

[9] E. W. Herold, "An analysis of the signal-to-noise ratio of ultra-high-frequency receivers," *RCA Rev.*, vol. 6, pp. 302–332, January, 1942.

$$F_{abc}=F_{ab}+(F_c-1)/G_{ab}=F_a+(F_b-1)/G_a+(F_c-1)/G_aG_b. \tag{16}$$

In most receivers the gains of the amplification stages are such that the noise figures of only the first two stages must be considered.

### Measurement of the Noise Figures of Two Networks

It may be desirable to determine $F_a$ by indirect measurements, particularly if $G_a$ is low. This may be done as follows. The noise figures $F_{ab}$ and $F_b$ may be measured by the method described for a single network. The gain $G_a$ may be obtained, by means of signal generators, as the increase in available signal power required to give a certain signal output reading when network $a$ is left out. The noise figure $F_a$ may then be calculated by means of (15).

A second method of noise-figure measurement includes the measurement of the ratio of the output noise of network $b$ with network $a$ in normal operation to that with network $a$ passive. This ratio is called the $Y$ figure and is especially useful in converter measurements. Network $a$ is said to be passive if its available output noise is only $KTB$ watts. In measuring $Y$ it is usually most convenient to replace the output impedance of network $a$ with an equal passive impedance at room temperature.

An expression for $F_a$ in terms of $Y$, $F_b$, and $G_a$ will be developed next. By definition,

$$Y = N_{ab}/N_b. \tag{17}$$

Formula (5) gives

$$F_b = (1/G_b)(N_b/KTB_b) \tag{18}$$

and

$$F_{ab} = (1/G_aG_b)(N_{ab}/KTB_{ab}). \tag{19}$$

The above three formulas give

$$F_{ab} = (F_bY/G_a)(B_b/B_{ab}). \tag{20}$$

It will also here be assumed that the two networks have the same ideal or square-top band-pass characteristic. Then

$$F_{ab} = F_bY/G_a. \tag{21}$$

Formulas (15) and (21) give

$$F_a = [F_b(Y-1)+1]/G_a. \tag{22}$$

Formula (22) is often simpler to use for the determination of $F_a$ than relationship (15) because it is easier to measure $Y$ than $F_{ab}$. Note also that $F_{ab}$ may be determined experimentally from the convenient relation given by (21).

### Conclusion

All signal and noise powers are in watts. It is confusing to use the decibel scale when noise powers are added, and it has not, therefore, been used in this paper.

In concluding, it is hoped that the definitions and symbols suggested will come into general usage. It should be pointed out that the paper is the result of a great many discussions during the past two years with scientific workers both inside and outside the Bell Telephone Laboratories.

# Grounded-Grid Radio-Frequency Voltage Amplifiers*

M. C. JONES†, ASSOCIATE, I.R.E.

***Summary***—**Triode radio-frequency amplifiers have come into extensive use for medium-high-frequency applications. The use of triodes results from the reduced noise-equivalent resistance of a triode amplifier as compared to a multigrid-type amplifier tube. It is not possible with a triode to use conventional circuits with the input into the grid circuit and the output from the plate circuit because this connection results in excessive output to input feedback which produces regeneration and even oscillation. The grounded-grid amplifier[1] circuit alleviates these difficulties by utilizing the grid as a shield between the input or cathode circuit and the output or plate circuit. Such a circuit exhibits certain peculiarities, particularly when several such stages are operated in tandem. Following is an analysis of the performance of several types of grounded-grid radio-frequency amplifiers.**

## I. Single Stage Amplifier

A SINGLE-STAGE amplifier consists of a high-impedance tuned circuit connected to the cathode of the amplifier tube and the load resistance connected in the plate. Such a circuit is shown in Fig. 1.

If we make the assumption that the impedance of the tuned circuit is high compared to the input resistance of the tube, it becomes evident that the signal current $I_1$ flowing to the cathode of the amplifier is the same as the plate current $I_1$ flowing into the plate of the amplifier. The tube then acts simply as a medium for transposing the input current, which may be from a low-impedance source, to the output circuit which may be a high-impedance source. From this we see that the gain of such a stage is the ratio of the output resistance ($R_L$) to the input resistance. The value of this input resistance will be determined later.

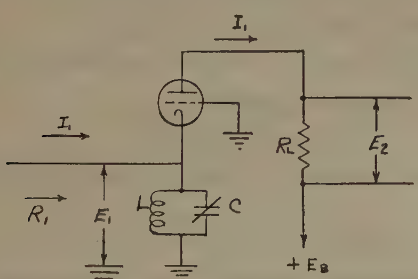


Fig. 1—Single stage grounded-grid amplifier.

A marked similarity exists between this type of circuit and the so called "cathode-follower" type of circuit. In the cathode follower the input and output voltages are identical to a first approximation, and in this circuit the input and output currents are identical. We might then coin the term "voltage-follower circuit" to apply to the cathode-follower and "current-follower circuit" to apply to the grounded-grid amplifier.

By inspection of Fig. 1 it is evident that

$$E_2 = I_1 R_L. \tag{1}$$

In any triode vacuum tube the fundamental frequency component of the signal current may be expressed by

$$I_1 = G_m(E_g + E_p/\mu). \tag{2}$$

Substituting the appropriate values for $E_g$ and $E_p$ in terms of $E_1$, $I_1$, and $R_L$ we have,

$$I_1 = G_m(-E_1 + (-I_1R_L - E_1)/\mu). \tag{3}$$

Solving for $E_1$, (3) becomes

$$E_1 = I_1((R_p + R_L)/(\mu + 1)). \tag{4}$$

We may now evaluate the input resistance to the amplifier tube as

$$R_1 = E_1/I_1 = (R_p + R_L)/(\mu + 1). \tag{5}$$

A curve showing $R_1$ as a function $R_L$ for a typical amplifier tube is shown in Fig. 2. This equation may be simplified, for the condition where $R_L \ll R_p$ and $\mu \gg 1$, to

$$R_1 = 1/G_m. \tag{6}$$

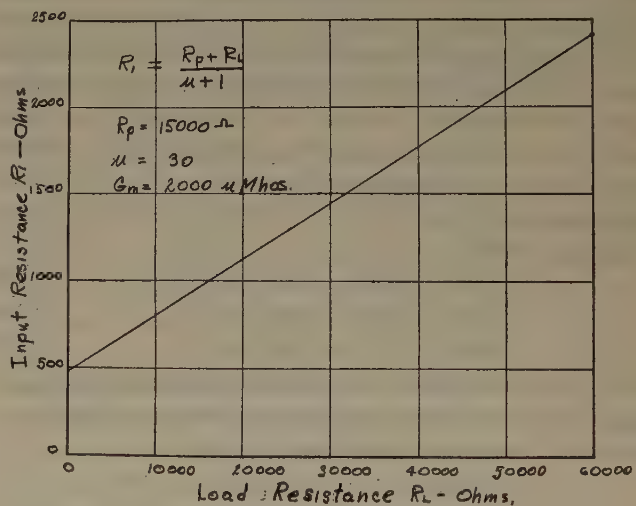


Fig. 2—Curve showing input resistance as a function of load resistance.

We may now evaluate the gain of the single amplifier by dividing (1) by (4).

$$\text{Gain} = E_2/E_1 = (R_L(\mu + 1))/(R_p + R_L). \tag{7}$$

This may be compared with the formula for the gain of an ordinary triode amplifier,

$$\text{gain} = R_L\mu/(R_p + R_L). \tag{8}$$

The slight increase in gain results from the fact that the output voltage is measured to ground which in the case of the grounded-grid amplifier is the grid circuit and in the case of the conventional amplifier is the cathode circuit.

If now we can make the assumption that $R_L \ll R_p$ and $\mu \gg 1$, equation (7) may be simplified to gain $= R_L G_m$.

* Decimal classification: R363.1. Original manuscript received by the Institute, September 14, 1943; revised manuscript received, March 9, 1944.
† Radio Corporation of America, RCA Victor Division, Camden, New Jersey.
[1] C. E. Strong, "The inverted amplifier," *Electronics*, p. 87; July, 1940.

## II. Tandem Circuits

It is customary to operate two or more grounded-grid amplifiers in tandem. In this case the load resistance ($R_L$) is some function of the input resistance of the following stage. A circuit showing three such stages terminated finally in the load resistance $R_L$ is shown in Fig. 3.

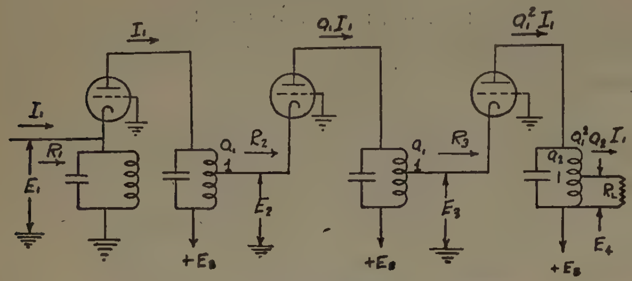

Fig. 3—Three grounded-grid amplifier circuits in tandem.

Before analyzing this circuit we must consider the coupling transformer which reflects the input resistance of each stage back to the load resistance of the preceding stage. For simplicity it is convenient to assume that this is a perfect transformer and that its output-to-input current ratio is the same as its input-to-output voltage ratio. If we define this ratio as $a$ it is then only necessary to multiply the input current by the appropriate values of $a$ through the various stages to determine the output current. In analyzing this circuit the most convenient method of approach is that of determining the input resistance of each stage as a function of the input resistance of the following stage. By reference to (5) we may write equations for the input resistances of the three stages as follows:

$$R_3 = (R_p + a_2^2 R_L)/(\mu + 1) \tag{9}$$

$$R_2 = (R_p + a_1^2 R_3)/(\mu + 1) \tag{10}$$

$$R_1 = (R_p + a_1^2 R_2)/(\mu + 1). \tag{11}$$

We may now solve for the input resistance $R_1$ as a function of the output resistance, the tube characteristics, and the transformer characteristics:

$$R_1 = R_p/(\mu + 1) + a_1^2 R_p/(\mu + 1)^2 + (a_1^4(R_p + a_2^2 R_L))/(\mu + 1)^3. \tag{12}$$

This equation may now be generalized for any number ($n$) of stages:

$$R_1 = \frac{R_p}{\mu + 1} + \frac{a_1^2 R_p}{(\mu + 1)^2} + \frac{a_1^4 R_p}{(\mu + 1)^3} + \cdots + \frac{a_1^{2(n-2)} R_p}{(\mu + 1)^{n-1}} + \frac{a_1^{2(n-1)}(R_p + a_2^2 R_L)}{(\mu + 1)^n}. \tag{13}$$

If the amplifier consists of a large number of stages such that the input resistance $R_1$ of the first stage is relatively independent of the output resistance $R_L$, or if the value of $R_L$ is selected such that the input resistance of each stage is identical, we then may evaluate what might be termed the "characteristic impedance" for this amplifier by setting $R_1 = R_2$ in (11).

$$R_1 = R_p/(\mu + 1 - a_1^2). \tag{14}$$

From this it may be seen that the characteristic input impedance of a grounded-grid amplifier is determined by the tube characteristics and by the coupling transformer. A curve showing the variation of input resistance with current ratio of the coupling transformer for a typical amplifier tube is shown in Fig. 4. From this curve it may be seen that the input resistance rises rapidly with values of $a_1$ approaching $\sqrt{\mu+1}$. We may now determine the over-all gain of $n$ grounded grid amplifier stages as follows:

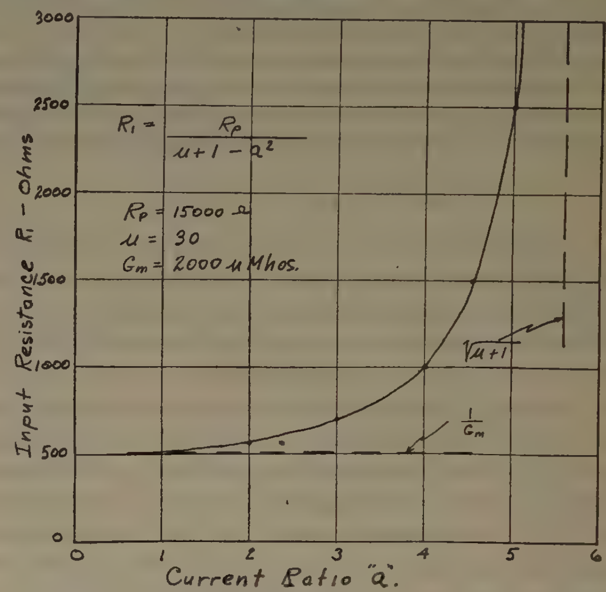


Fig. 4—Curve showing input resistance of an amplifier terminated in its "characteristic impedance" as a function of $a$.

$$\text{Output current } (I_n) = a_1^{n-1} a_2 I_1 \tag{15}$$

$$\text{Output voltage } (E_n) = I_n R_L = I_1 R_L a_1^{n-1} a_2. \tag{16}$$

The input voltage $E_1$ is given by

$$E_1 = I_1 R_1. \tag{17}$$

The over-all gain may be determined by dividing (16) by (17).

$$\text{Over-all gain} = (R_L a_1^{n-1} a_2)/R_1 \tag{18}$$

or substituting the value of $R_1$ for $n$ stages given in (13) we have,

$$\text{over-all gain} = \frac{R_L a_1^{n-1} a_2}{R_p/(\mu + 1) + a_1^2 R_p/(\mu + 1)^2 + \cdots + a_1^{2(n-2)} R_p/(\mu + 1)^{n-1} + a_1^{2(n-1)}(R_p + a_2^2 R_L)/(\mu + 1)^n}. \tag{19}$$

If the stages are designed so that they exhibit the "characteristic input impedance" or if a large number of stages is used such that the input resistance is relatively independent of the output resistance $R_L$, equation (19) may be simplified to

$$\text{over-all gain} = (R_L a_1^{(n-2)} a_2(\mu + 1 - a_1^2))/R_p. \tag{20}$$

If $\mu \gg (1 - a_1^2)$, equation (20) may be simplified to

$$\text{over-all gain} = G_m R_L a_1^{n-1} a_2. \tag{21}$$

From this equation it is evident that the gain of several

grounded-grid amplifier stages in tandem is a function only of the current ratio of the coupling transformers, the $G_m$ of any one tube (assuming all tubes have the same value of $G_m$), and the load resistance $R_L$. It will be shown later an optimum value of $a$ exists if we consider the input loading due to transit-time effect.

## III. $\pi$-Section Tandem Circuits

It is sometimes more convenient practically to design the coupling transformers as a $\pi$-section filter in the manner shown in Fig. 5.

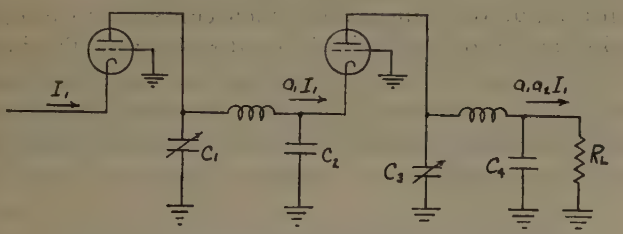


Fig. 5—$\pi$-section coupling transformers in two-stage grounded-grid amplifier.

It can be shown that the current ratio of these coupling transformers to a fair degree of approximation is given by

$$a_1 = C_2/C_1 \tag{22}$$

$$a_2 = C_4/C_3. \tag{23}$$

From this it may be seen that for large gains $C_2$ should be greater than $C_1$, and $C_4$ should be greater than $C_3$. To obtain the maximum range from a given capacitance change in the tuning capacitor it is desirable then to make $C_1$ and $C_3$ the variable elements and to use a fixed capacitor for $C_2$ and $C_4$.

With the above definitions of $a_1$ the previous equations for over-all gain (18), (19), (20), (21), apply with equal accuracy to this circuit.

## IV. Antenna Input Circuits

Most amplifiers of this type operate from a low-impedance antenna circuit and feed into a constant impedance, usually the input impedance of a converter. A typical antenna circuit feeding directly into the input resistance of a grounded-grid amplifier is shown in Fig. 6.

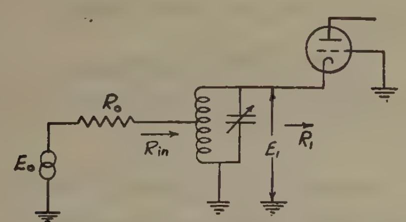


Fig. 6—Antenna input circuit for grounded-grid amplifier.

If we assume that the loss in the input transformer is negligible and that the input circuit matches the antenna resistance it can be shown that the gain of the antenna circuit is given by

$$\text{gain} = G_1 = E_1/E_0 = 1/2\sqrt{R_1/R_0}. \tag{24}$$

The over-all gain for $n$ stages plus the above antenna circuit is then given by

$$\text{over-all gain} = E_n/E_0 = ((R_La_1^{n-1}a_2)/R_1)/2\sqrt{R_1/R_0}. \tag{25}$$

Simplifying the above equation

$$\text{over-all gain} = 1/2(R_La_1^{n-1}a_2)/2\sqrt{R_1R_0}. \tag{26}$$

In this equation we may use values of $R_1$ as determined previously in (13) or (14). If the following approximation is valid,

$$R_1 = 1/G_m. \tag{27}$$

The formula for over-all gain simplifies to

$$\text{over-all gain} = 1/2R_La_1^{n-1}a_2\sqrt{G_m/R_0}. \tag{28}$$

From this it may be seen that the over-all voltage gain depends entirely on the transformer design, the input and output impedances, and the $G_m$ of the amplifier tubes. An increase of the $G_m$ of all the tubes produces an increase in over-all gain of only the square root of the $G_m$ change of any one stage regardless of the number of stages. Consider a typical two-stage grounded-grid amplifier feeding from an antenna to the input of a converter with the following characteristics:

$G_m$ = 2000 micromhos $\quad a_1 = 3$
$R_0$ = 50 ohms $\quad a_2 = 2$
$R_L$ = 1000 ohms

The over-all gain as calculated by (28) then becomes

$$\text{over-all gain} = 1/2\times1000\times3\times2\sqrt{(2000\times10^{-6})/50} = 19.$$

## V. Noise Considerations

The performance of any receiver is limited by noise voltages.[2,3] This noise is of two distinct types: Thermal-agitation noise, or noise developed within the circuits, and shot-effect noise; or noise developed in the plate of the amplifier tubes. The thermal-agitation noise over any band of frequency can be predicated accurately if the value of circuit resistance is known. The shot-effect noise is more difficult to determine. In general this noise is referred to the grid circuit and expressed as a "noise-equivalent resistance" for the vacuum tube used, and calculations may then be made in the same manner as with thermal-agitation noise.

The noise voltage generated in series with any resistor $R$ as measured by an amplifier having a bandwidth $B.W.$ is given by the expression:

$$E_n^2 = 4KTR\ B.W. \tag{29}$$

where

$K$ = Boltzmann's constant = $1.37\times10^{-23}$ watt second per degree
$T$ = temperature in degrees Kelvin
$R$ = resistance in ohms
$B.W.$ = bandwidth of the amplifier in cycles per second

Since most of our calculations are based on normal or

[2] F. B. Llewellyn, "A rapid method of estimating the signal-to-noise ratio of a high gain receiver," Proc. I.R.E., vol. 19, pp. 416–420; March, 1931.
[3] E. W. Herold, "An analysis of the signal-to-noise ratio of ultra-high-frequency receivers," *RCA Rev.*, vol. 6; January, 1942.

room temperature, it is possible to simplify (29) as follows:

$$E_n^2 = 1.66 \times 10^{-20} \times B.W. \times R \tag{30}$$

where $E_n$ = the root-mean-square noise voltage in series with the resistance $R$ at room temperature, 30 degrees centigrade, or 303 degrees Kelvin.

*Measurement of Noise Factor (NF) by the 4KT Method*

The noise factor of a receiver expresses its ability to receive and detect weak signals in the presence of noise generated both in the antenna resistance and the receiver itself. The noise factor may be defined as the ratio of the signal-to-noise ratio of a perfect receiver, to the signal-to-noise ratio of the receiver under test (both measured under the same conditions). A perfect receiver is defined as one with infinite input impedance, and no internal noise. Expressing this in an equation we have

$$NF = \frac{(E_s/E_n)}{(E_s/E_n^1)} = \frac{E_n^1}{E_n} \tag{31}$$

where,

$E_s$ = signal voltage used to measure both receivers

$E_n$ = equivalent-noise voltage of the perfect receiver referred to the input circuit

$E_n^1$ = equivalent-noise voltage of the receiver under test referred to the input circuit

From this it is apparent that the noise factor is the ratio of two noise voltages, one of which may be measured and the other calculated.

In order to measure the equivalent-noise voltage of a receiver, one would proceed as follows:

1. Connect a signal generator, of negligible internal impedance, to the receiver input terminals in series with a resistance equal to the input resistance of the receiver. It should be noted here that the $4KT$ method applies only when the internal impedance of the signal generator is small.
2. Connect a root-mean-square voltmeter to the output of the intermediate-frequency amplifier of the receiver. A direct-current voltmeter connected across the load resistor of second detector may be used with only slight error.
3. With the signal generator output reduced to zero, adjust the gain control until the noise voltage as read on the root-mean-square meter is some convenient value, say 1 volt.
4. Adjust the input voltage from the signal generator until the noise voltage has increased to $\sqrt{2}$ times its original value.

The voltage output from the signal generator is now equal to the equivalent-noise voltage of the receiver referred to the input circuit; or the voltage $E_n^1$ in (31).

The equivalent-noise voltage ($E_n$) of a perfect receiver, with characteristics similar to the receiver under test, is the noise voltage generated by the antenna resistance. This value of resistance is usually equal to the input resistance of the receiver under test. From (30) this voltage is

$$E_n = 1.29 \times 10^{-10}\sqrt{B.W. \times R_0} \tag{32}$$

where

$B.W.$ = bandwidth of receiver under test in cycles per second

$R_0$ = antenna series resistance in ohms

$E_n$ = equivalent-noise voltage of the perfect receiver

The noise factor may now be determined as the ratio $E_n^1/E_n$.

Since it is not possible to obtain a perfect receiver with infinite input impedance, it is useful to consider what might be called a "practically perfect" receiver with a finite input resistance, but which contains no internal noise sources other than this input resistance. The input circuit of such a receiver connected to a source of signal voltage $E_0$ is shown in Fig. 7 (a). The noise sources and their equivalent noise voltage are shown in Fig. 7(b), and the previous perfect receiver is shown in Fig. 7 (c).

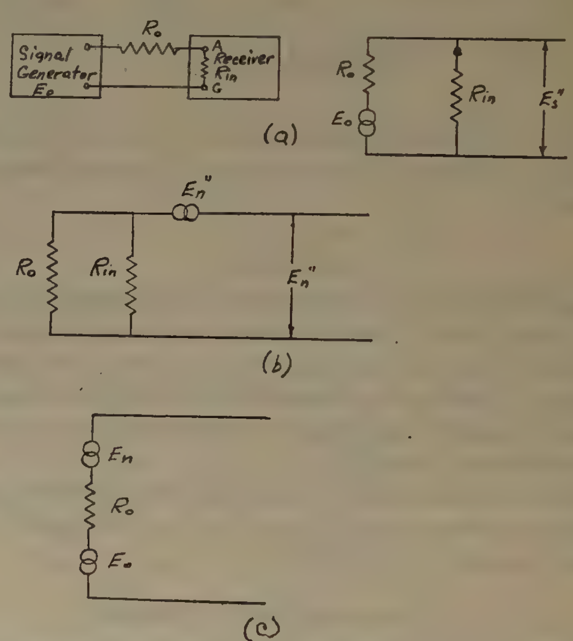


Fig. 7—Receiver input circuit showing noise and signal voltages. 4KT method of noise-factor measurement.
(a) Input circuit showing signal voltage
(b) Input circuit showing noise voltage
(c) Perfect receiver

Solving for the receiver input voltage $E_s''$ from Fig. 7 (a) we have,

$$E_s'' = E_0(R_{in}/(R_0 + R_{in})). \tag{33}$$

The noise voltage $E_n''$ from Fig. 7 (b) is given by the expression

$$E_n'' = 1.29 \times 10^{-10}\sqrt{B.W.(R_0R_{in}/(R_0 + R_{in}))}. \tag{34}$$

Performing these same operations on the perfect receiver shown in Fig. 7 (c) we have,

$$E_s = E_0 \tag{35}$$

$$E_n = 1.29 \times 10^{-10}\sqrt{B.W. \times R_0}. \tag{36}$$

If we now consider the practically perfect receiver as one under test we have, from (31),

$$NF=\frac{(E_s/E_n)}{(E_s''/L_n'')}=\frac{(E_0/(1.29\times10^{-10}\sqrt{B.W.\times R_0}))}{\dfrac{E_0(R_{in}/(R_0+R_{in}))}{1.29\times10^{-10}\sqrt{B.W.\times R_0R_{in}/(R_0+R_{in})}}}$$

$$=\sqrt{\frac{R_0+R_{in}}{R_{in}}}=\sqrt{\frac{1+R_{in}/R_0}{R_{in}/R_0}}. \quad (37)$$

The curve in Fig. 8 shows the noise factor as a function of the ratio $R_{in}/R_0$. From this it may be seen that the noise factor of a matched receiver cannot exceed $\sqrt{2}$, or 3 decibels from thermal noise, even though there are no noise sources within the receiver except the input resistance. A slight advantage may be gained by mismatching the receiver in the direction of making $R_{in}$ greater than $R_0$. This advantage is more theoretical than practical, however, since mismatch always introduces additional losses due to reflections on the transmission line connecting the receiver with the antenna. It also tends to make the performance of the receiver a function of the transmission-line length since the transmission line is then a resonant circuit. At lower frequencies where the transmission line is short compared with a quarter wavelength, some advantage may be gained by mismatching the antenna circuit.

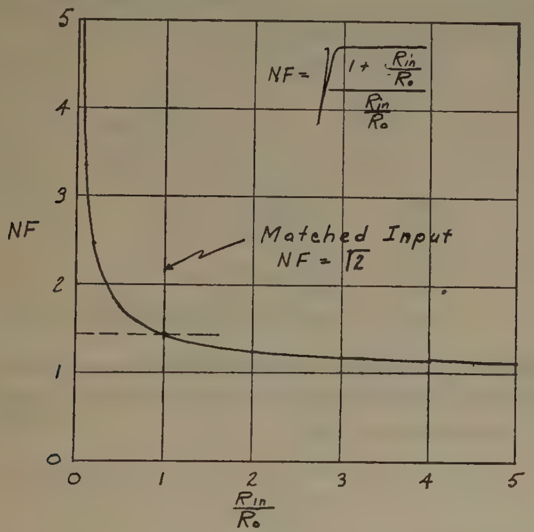


Fig. 8—Curve showing the noise factor as a function of receiver input resistance, for a receiver with no internal noise ($4KT$ basis).

### *Measurement of Noise Factor by the 1KT method*

In the preceding analysis the signal-generator voltage ($E_0$) was connected in series with the antenna resistance ($R_0$), and as a result, the thermal-agitation noise was calculated on the so-called $4KT$ basis. In some signal generators, however, the internal resistance of the generator itself is equal to the antenna resistance, and the voltage as read by the signal generator is then equal to the voltage impressed on the receiver input terminals. A schematic diagram of such an arrangement, together with its equivalent circuits, is shown in Fig. 9.

With this arrangement the noise voltage must be re-evaluated so that it compares with the readings of voltage from the signal generator. The signal-generator voltage reading is now half its value in the preceding connection shown in Fig. 7 (a) and the noise voltage must be reduced by a factor of 2 for a direct comparison with it. This gives rise to an expression for noise voltage which is not strictly accurate, but which gives proper results when compared with the equivalent-noise voltage of the receiver shown in Fig. 9 measured by the method previously outlined. This expression for noise voltage is

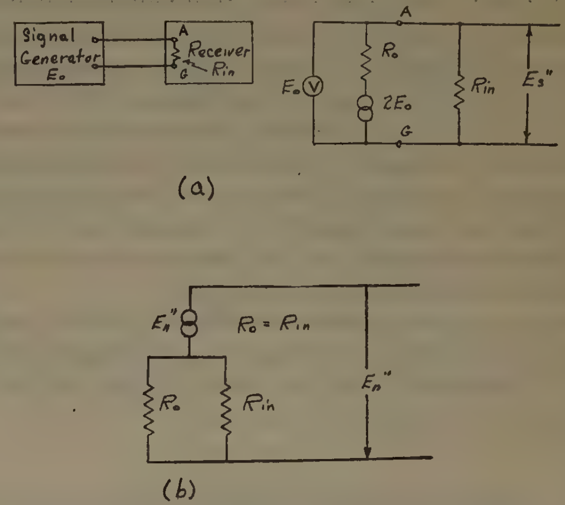


Fig. 9—Receiver input circuit with matched impedance signal generator. $1KT$ method of noise-factor measurements.
(a) Input circuit showing signal voltage
(b) Input circuit showing noise voltage

$$(E_n'')^2 = 1KTR_0\,B.W. \quad (38)$$

At $T=303$ degrees Kelvin (30 degrees centigrade) this becomes

$$(E_n'')^2 = 0.415\times10^{-20}\times B.W.\times R_0 \quad (39)$$

or

$$E_n'' = 0.645\times10^{-10}\sqrt{B.W.\times R_0}. \quad (40)$$

This method of measuring noise factor is based on the so-called $1KT$ noise level. For any given receiver the noise factor as determined by either of these methods should be identical. The method used depends on the internal resistance of the signal generator. In actual practice, the method used may quite often be a compromise between the $1KT$ and the $4KT$ methods. This condition occurs when the internal resistance of the signal generator is less than the input resistance of the receiver, but is still not negligible. By any method of measurement, however, a matched receiver with no internal noise has a theoretical maximum performance of 3 decibels from thermal noise.

### *Calculation of Receiver Noise and Noise Factor*

The noise generated in the plate of a vacuum tube may be conveniently expressed as the noise-equivalent resistance of the tube referred to the grid circuit. This is considered to be the resistance connected in series with the grid of a perfect amplifier tube having the same gain as the tube under consideration and generating the noise normally generated in the plate of the tube.

If one makes the assumption that the circuit noise in a radio-frequency amplifier is negligible compared with

the noise generated in the antenna resistance, the input resistance of the receiver, and the shot-effect noise of the radio-frequency amplifier tubes, one may calculate the noise factor of the radio-frequency amplifier from a knowledge of its characteristics.

The noise-equivalent resistance of the first stage may be referred to the input circuit by dividing its value by the square of the gain of the antenna circuit. If additional stages, other than the first stage, contribute to the noise-equivalent voltage of the receiver, they may be taken into account by dividing the noise-equivalent resistance of each stage by the square of the total gain from the input circuit to the grid of the stage. Fig. 10 represents the input circuit to a receiver and the noise equivalent resistances of the first and second radio-frequency amplifier tubes.

Since the noise-equivalent resistances and the parallel equivalent of the input and antenna resistance may be added numerically, the total noise voltage referred to the input of the amplifier is given by

$$E_n' = 1.29 \times 10^{-10}\sqrt{B.W.\ [R_0R_{in}/(R_0 + R_{in}) + R_1'/G_0^2 + R_2'/G_0^2G_2^2]} \quad (41)$$

where

$R_1'$ = noise-equivalent resistance of first radio-frequency amplifier tube
$R_2'$ = noise-equivalent resistance of second radio-frequency amplifier tube
$G_0$ = antenna circuit gain measured from receiver input terminals to input of the first radio-frequency amplifier tube
$G_2$ = gain of first radio-frequency amplifier stage.

The signal voltage with which this noise voltage should be compared is given by (33).

$$E_s' = E_0(R_{in}/(R_0 + R_{in})). \quad (33)$$

The signal-to-noise ratio of this receiver is then,

$$\frac{E_s'}{E_n^1} = \frac{E_0(R_{in}/(R_0 + R_{in}))}{1.29 \times 10^{-10}\sqrt{B.W.\ [(R_0R_{in}/(R_0 + R_{in})) + R_1'\ G_0^2 + R_2'/G_0^2G_2^2]}}. \quad (42)$$

The signal-to-noise ratio of a perfect receiver, as shown in Fig. 7 (c) is,

$$\frac{E_s}{E_n} = \frac{E_0}{1.29 \times 10^{-10}\sqrt{B.W.\ R_0}}. \quad (43)$$

The noise factor of the receiver is obtained by dividing (43) by (42).

$$NF = \frac{(E_s/E_n)}{(E_s'/E_n')} = \frac{1/\sqrt{R_0}}{\dfrac{(R_{in}/(R_0+R_{in}))}{\sqrt{R_0R_{in}/(R_0+R_{in}) + R_1'/G_0^2 + R_2'/G_0^2G_2^2}}}$$

$$= \frac{R_0+R_{in}}{R_{in}}\sqrt{\frac{R_{in}}{R_0+R_{in}} + \frac{R_1'}{R_0G_0^2} + \frac{R_2'}{R_0G_0^2G_2^2}}. \quad (44)$$

This equation simplifies to (37) when $R_1'$, $R_2'$, etc., $=0$.

It should be noted that the value for $G_0$, the antenna circuit gain, is now different from the value $G_1$ previously calculated in (24). Since our reference point for

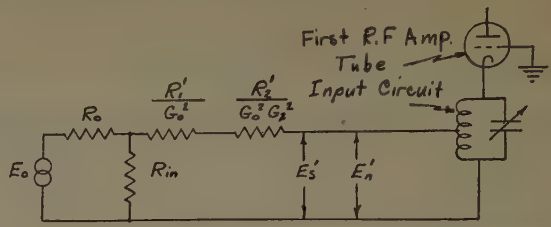


Fig. 10—Noise sources within a receiver.

signal-to-noise calculations is the input circuit of the receiver, the gain must be measured from the input terminals to the grid or cathode of the first tube as shown in Fig. 11.

This gain ($G_0$) is greater than the gain ($G_1$) calculated by (24) and is given by

$$G_0 = \text{Gain} = E_1/E_0' = \sqrt{R_1/R_0}. \quad (45)$$

Since the antenna-gain equations (24) and (45) assume matched-input conditions,

$$G_0 = 2G_1. \quad (46)$$

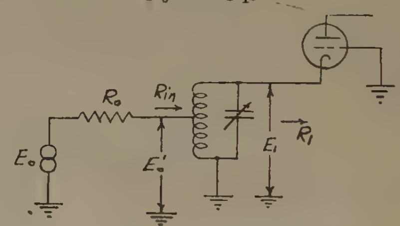


Fig. 11—Antenna input circuit.

Simplifying (44) for matched-input conditions and substituting the generally accepted value for antenna gain $G_1$, as given by (24),

$$NF = \sqrt{2 + R_1'/Z_0G_1^2 + R_2'/Z_0G_1^2G_2^2 + \cdots}. \quad (47)$$

Consider now a grounded-grid amplifier circuit with several stages having the following characteristics:

$$R_0 = 50 \text{ ohms}$$
$$R_{in} = 50 \text{ ohms.}$$

Input resistance of first tube = 500 ohms

$$G_1 = 1/2\sqrt{500/50} = 1.58$$
$$G_2, G_3, G_4, \cdots = 3(a = 3.3)$$
$$R_1', R_2', R_3', \cdots = 300 \text{ ohms.}$$

Calculating the noise factor by (45) we have,

$$NF = \sqrt{2 + 2.4 + 0.27 + 0.03 + \cdots} \quad (48)$$
$$= 7 \text{ decibels from thermal noise.}$$

From this it may be seen that the principal source of

noise in this amplifier is the shot-effect noise of the first radio-frequency amplifier tube. The following stages contribute a progressively smaller amount of noise and the noise contributed by the third stage is insignificant. The converter might take the place of the third stage. If we assume the noise equivalent resistance of the converter to be 1000 ohms and recalculate the noise factor for a two-stage radio-frequency amplifier followed by the converter, we obtain the following results:

$$NF = \sqrt{2 + 2.4 + 0.27 + 0.1}$$
$$= 7 \text{ decibels from thermal noise.} \qquad (49)$$

## VI. Coupling Transformer Designs and Optimum Values for $R_L$

From the previous analysis it appears that the gain increases without limit as the value of $a$ for the coupling transformer increases. In practice, certain physical limitations produce an optimum value of $a$ which should be used in any grounded-grid amplifier to obtain maximum gain. The most serious limitation is that imposed by a component of input resistance resulting from the transit-time effect.[4] Fig. 12 shows a single-stage amplifier with this component of input resistance ($R_x$) in parallel with the normal input resistance $R_1$.

It is convenient to analyze this amplifier on the basis of power rather than voltage gain since the cathode and plate current are common. This power gain then becomes the ratio of output to total input resistance. Solving for power gain for the circuit shown in Fig. 12 we have

$$W_{in} = I_1^2 R_1 (R_1 + R_x)/R_x \qquad (50)$$
$$W_{out} = I_1^2 R_L. \qquad (51)$$
$$\text{power gain} = R_1 R_x/(R_1 + R_1 R_x). \qquad (52)$$

For any single-stage amplifier,

$$R_1 = (R_p + R_L)/(\mu + 1). \qquad (53)$$

Substituting this value of $R_1$ in (52) and simplifying we have

$$\text{power gain} = \frac{R_1 R_x(\mu + 1)^2}{R_L^2 + R_L[2R_p^2 + R_x(\mu + 1)] + R_p^2 + R_x R_p(\mu + 1)}. \qquad (54)$$

By taking the derivative of this expression and setting it equal to zero we may determine the value for $R_L$ which produces the maximum power gain. Since we may design a transformer with fairly high efficiency to match the output of this stage into input of another stage this maximum power gain then corresponds to a maximum voltage gain into any constant impedance. This optimum value for $R_L$ is,

$$R_L = R_p \sqrt{1 + (R_x(\mu + 1)/R_p)}. \qquad (55)$$

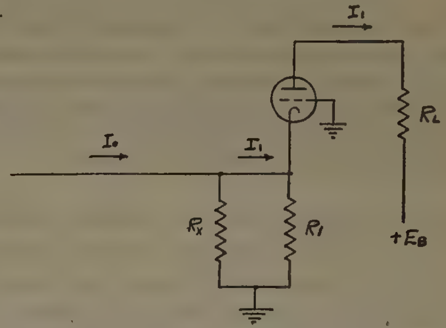


Fig. 12—Single-stage grounded-grid amplifier considering transit-time effect.

For conditions where $\mu$ is considerably greater than 1, (55) simplifies to

$$R_L = R_p \sqrt{1 + R_x G_m}. \qquad (56)$$

If we now consider the coupling transformer between this stage and the next stage whose input resistance is $R_2$ we have the expression

$$R_L = a^2 R_2. \qquad (57)$$

From this we may determine an optimum value for $a$ as follows:

$$a^2 = R_p/R_2 \sqrt{1 + R_x G_m}. \qquad (58)$$

If the input resistance of the next stage is the "characteristic impedance" as given by (14) the value for $a$ may be written in the form

$$a^2 = ((\mu + 1)\sqrt{1 + G_m G_x}/(1 + 1\sqrt{+ G_m R_x})). \qquad (59)$$

At extremely high frequencies the value of $R_x$ decreases rapidly. If the value of $G_m R_x$ is considerably less than unity, (59) simplifies to

$$a = \sqrt{\mu/2}.$$

For the amplifier considered in Fig. 2, the value for $a$ under this condition is approximately 3.9. In practice other limitations are present such that a value slightly less than this is often used. For normal operating conditions the value for $a$ of approximately 3 represents the practical optimum.

[4] W. R. Ferris, "Input resistance of vacuum tubes as ultra-high-frequency amplifiers," Proc. I.R.E., vol. 24, pp. 82–105; January, 1936.

# Joint Army-Navy Tube Standardization Program*

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*Summary*—**The steps taken by the Bureau of Ships and the Signal Corps to effect standardization of radio tubes are recounted briefly. The procedures for handling tube problems, including the JAN-1 Tube Subcommittee, its duties and reasons therefor, are described. Tube specifications, preferred lists, special selection, and type approvals are among the subjects discussed.**

THIS war depends to an amazing extent upon electronics. The many advances in radio and related equipments now in use by our Armed Forces would not have been possible without electronic tubes. It is not desired to minimize the importance of circuit and related developments, but without tubes, these circuits and developments would not be possible.

Consequently, it was evident at an early stage that careful attention should be given to electron-tube matters in order to secure best results. Before entering the conflict, the Army and Navy each had its own systems of tube nomenclature, which were unrelated to the Radio Manufacturers Association and other commercial type numbers. Radio engineers, hurriedly called into government service, had to contend with these unfamiliar Army and Navy numbers which were so unsuited to the desirable qualities of interchangeability, common stockpiles, joint inspection, and general efficiency.

The Navy was the first to do away with its special nomenclature for tubes by adopting the use of RMA and commercial type numbers in 1940.

Early in 1942 work was begun to prepare a joint Army-Navy specification for tubes which would be based upon the use of RMA and commercial type numbers, and the Army decided to do away with the special nomenclature which it had been using.

Through the use of a single specification for both major branches of the Armed Services, it was felt that greatly improved results could be obtained by the elimination of more than one specification per tube type by the easy interchangeability, which would result between Army and Navy tubes, and by the possibility of creating joint stockpiles. Furthermore, it could save manpower by eliminating dual inspection at a manufacturer's plant, since, because of their familiarity with this common specification, either Army or Navy inspectors could inspect for either branch of the service.

It was, at first, planned to prepare these joint specifications by combining the previous Army and Navy specifications for each type. Thus if the vacuum-tube specification for a particular tube called for a plate-current range of from 4 to 8 milliamperes, and the former Navy specification specified from 5 to 9 milliamperes, the joint specification would require 5 to 8 milliamperes. This plan was later dropped in favor of basing each specification on actual needs of the Services, as well as upon good engineering practice and production capabilities.

Naturally, in a venture of this sort, the first specification which resulted, after several months' work by representatives of the Navy and the Signal Corps, was not perfect. The JAN-1 Specification for Tubes, Radio Electron, as it was named, was put into use by the Signal Corps and Bureau of Ships in March, 1943, but caused such a multitude of objections and suggested changes, due to insufficient test equipment and lack of data in the tube manufacturer's plant to enable him to know how closely his product could be controlled, that the granting of waivers to help manufacturers through this transition period was authorized. At the same time the tube manufacturers were requested to work together to recommend changes in the specification which would make it more acceptable to them. This joint action was necessary to insure that the Services had but one co-ordinated specification suggestion from Industry to consider for each type. Through the able assistance of the RMA, the various tube manufacturers assembled their comments and suggestions, after a considerable amount of time spent by the representatives of these manufacturers in meetings of the various RMA Tube Committees. These suggestions were then conveyed to the Army and Navy engineers working on the specification and several joint Industry-Service meetings took place for further discussion of uncertain or questionable points. By the latter part of 1943 a revision of the original specification was completed and issued, entitled Joint Army-Navy Specification JAN-1A for Radio Electron Tubes. This specification is now mandatory for all Signal Corps and Navy tube contracts.

There are quite a few tube types for which JAN Specifications have not yet been written, but these are being completed as promptly as possible. With but few exceptions, all but some of the special purpose and classified high-frequency types are not yet specified, but procurement of these is being made on temporary Army or Navy specifications, written to conform to the JAN style.

In a brief résumé, the preceding remarks have brought us up to the present. Now the working organization, its duties, and aims will be described.

Joint Army-Navy specifications, of which the Radio Electron Tube Specification is the first, are controlled and authorized by a JAN Committee comprised of members of the Army Service Forces and Naval Office of Procurement and Materiel. This committee has authorized a subcommittee to handle the tube specification and has specified its duties. It is known as the JAN-1

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‡ United States Navy, Bureau of Ships, Washington, D. C.

Tube Subcommittee. Its members are representatives of the various Service organizations concerned and are as follows:

1. The Navy—Bureau of Ships.
2. The Signal Corps Aircraft Signal Agency, a division of which is well known as the Aircraft Radio Laboratory.
3. The Signal Corps Ground Signal Agency, comprising Camp Evans Signal Laboratory and the Laboratories in and near Fort Monmouth.
4. The Signal Corps Standards Agency, which is essentially a branch of the Office of the Chief Signal Officer.

At this time the actual representatives are:

Mr. J. W. Greer for the Navy
Lieutenant E. A. Anderson for the Signal Corps Ground Signal Agency
Captain F. D. Langstroth for the Signal Corps Standards Agency, chairman
Captain C. W. Martel for the Signal Corps Aircraft Signal Agency

All have had technical experience in the radio-tube industry. Other personnel from any of these member organizations may be brought in by the regular representative if needed.

The Signal Corps Standards Agency acts as the secretariat for the Army members of the JAN-1 Tube Subcommittee, as well as co-ordinating and administrative center for Army matters that come under the jurisdiction of this subcommittee. The JAN-1 Tube Subcommittee meets regularly twice each month, but may be called together by the Signal Corps Standards Agency or the Bureau of Ships whenever required. For instance, special meetings have been held to discuss with various manufacturers, suggestions which had been made in connection with certain types which they produce. If a particularly urgent matter arises, the subcommittee members may be asked for a decision by telephone or telegraph, so that there is no time lost in securing an answer to important problems.

The official duties of the JAN-1 Tube Subcommittee as supplied by the JAN Committee are as follows:

1. To prepare and establish standard tube specifications and ratings.
2. To review, revise, and correct these specifications as required by technical improvements, changes in Service requirements, etc.
3. To co-ordinate tube-specification matters with the Services and Industry.
4. To supervise the distribution of the specifications to all concerned.

In addition to these duties which apply directly to the JAN-1A Specification, the members of the subcommittee have additional duties as follows:

1. The consideration of proposed tests to determine if the resulting tubes are equivalent to those which meet tests required by the JAN-1A specification.
2. The issuance of tube type approvals under the JAN-1A Specification after qualification tests have been made.
3. The compiling and revising of the Army-Navy preferred lists of vacuum tubes.
4. The recommending of security classifications for vacuum tubes.
5. Co-ordination with Canadian Armed Services in their use of the JAN-1A Specification.
6. Co-ordination with the RMA on tube numbering and test methods.

It may now be well to take up some of these duties in detail in order to indicate their scope and method of handling.

First it is desirable to explain the method by which these electron-tube matters are referred to the JAN-1 Tube Subcommittee. Any electron-tube manufacturer, equipment manufacturer, Service laboratory, or other government organization concerned with JAN tubes may originate a matter requiring action by the subcommittee. This matter should then be sent to the Signal Corps Standards Agency or Bureau of Ships depending upon the Service branch most concerned. Army contractors and subcontractors will follow the usual procedure of submitting their problems through the prime, or equipment, contractor to the contracting officer or laboratory concerned, which will forward them to the Signal Corps Standards Agency with their recommendations. Navy contractors may contact the Bureau of Ships, Code 930A, directly. The Bureau of Ships will then pass these subjects on to the Signal Corps Standards Agency where all such matters will be put on the agenda of the next subcommittee meeting unless immediate action is requested. After action by the JAN-1 Tube Subcommittee, its decision on the matter will be transmitted by either the Signal Corps Standards Agency or Bureau of Ships to the organization which made the original request.

The IRE Standards on Electronics of 1938 are used, where applicable, for references in the JAN-1A specification and it is hoped that these standards will be brought up to date soon to include all of the tube characteristics for which present military equipments have created a need for standardization. The results of the work of the recently organized Subcommittee on Advanced Developments of the IRE are awaited with interest by the JAN-1 Tube Subcommittee.

The duty of preparing and establishing JAN tube specifications is carried out in co-operation with the tube manufacturers, Service laboratories, and other organizations concerned. Requirements which the tube must meet in order that it operate satisfactorily in all equipments are compiled and compared with tests and test limits suggested by the manufacturers. The final specification as adopted combines these two viewpoints into a set of tests and limits which come as closely as possible to satisfying equipment requirements without causing the manufacturer undue hardship through

excessive rejections, complicated production or test equipment, or similar factors which would adversely affect production. In some cases several conferences between equipment engineers, JAN-1 Subcommittee members, and tube manufacturers' representatives may be necessary before a specification is mutually satisfactory.

Naturally all concerned must have a say in what tests will be required for each tube. Any one service organization or tube or equipment manufacturer cannot alone write a specification which will be sure to be suitable for all users of the tube. It may seem reasonable that the services which use the tube should write the specifications to insure that the tubes will then meet the Service needs, but this is not satisfactory, for the tube manufacturers must also have an opportunity to comment, in order to insure that the specification can be met through its being in accordance with good production and engineering design and practice, even though this may not allow the Services to make as "tight" a specification as they might prefer. In other words, it is sometimes necessary to sacrifice the ultimate in desired performance in order that the tube can be produced with reasonable efficiency and by more than one tube manufacturer. After this complete co-ordination with all concerned, the resulting specification as written, insures the best possible operation of the tube in the equipments using it, without including requirements which would cause excessive production difficulties.

In securing co-ordination with the tube manufacturers on problems of a general nature which affect all, the JAN-1 Tube Subcommittee has worked, and will continue to work, with the various RMA tube committees. These RMA groups are made up of engineers from the tube manufacturers and comprise committees on receiving tubes, transmitting tubes, cathode-ray tubes, etc. Matters which arise are referred to the appropriate RMA committee through the Electron Section of the RMA Engineering Department. The recommendations of the RMA committee are then considered by the JAN-1 Tube Subcommittee and adopted as completely as possible consistent with Service requirements. Furthermore, some of the RMA tube committees are carrying on investigations requiring appreciable time and data gathering, with a view to recommending improved tests for and information regarding the use of tubes under less usual operating conditions.

Tube specifications do not necessarily remain fixed after their initial acceptance by the JAN-1 Tube Subcommittee, because improved design and production techniques, and changed equipment requirements may involve specification changes. Thus the subcommittee's duties include that of reviewing and correcting specifications as necessary. If errors are found in the JAN-1A specification or if it is desired to recommend changes, the information should be sent either to the Signal Corps Standards Agency or to the Bureau of Ships, Code 930A. This will insure that they are referred to the JAN-1 Subcommittee for consideration. In submitting recommended changes it is necessary that complete reasons for the recommendations together with substantiating data be included.

From the preceding description, it will be seen that the specification is handled in a manner which enables it to reflect the latest in government and industry advancements and requirements.

The JAN-1A Specification does not bear any security classification since specifications for confidential types are issued individually to those organizations which have need for them. Such specifications must be kept in accordance with government security regulations.

The JAN-1A Specification is available to any contractor or government organization which requires it. The distribution of the JAN-1A Specification is not, however, unlimited, due to practical considerations such as preparation and printing time, distribution problems, and cost. In order to insure that the specification is properly distributed, all requests for copies must be approved by the JAN-1 Tube Subcommittee. Requests for copies may be sent either to the Signal Corps Standards Agency or the Bureau of Ships.

It is well to point out that the JAN-1A Specification is still incomplete in that all tube characteristics are not necessarily controlled by the required tests. Therefore, any tube application which requires the control of less frequently used tube characteristics should be checked against the specification to insure that it provides tests to control those characteristics. For example, the following tube characteristics are not usually controlled by the specification:

1. Triode and suppressor cutoff characteristics of pentodes.
2. Plate-current cutoff tolerances for types used as resistance-capacitance oscillators.
3. Radio-frequency characteristics of types originally designed for audio-frequency circuits.

If a particular equipment requires the control of characteristics not now controlled, the matter should be brought before the JAN-1 Tube Subcommittee together with data to enable the formulation of a suitable test. The subcommittee will then do all possible to arrange for specification revisions to accomplish the desired result.

At this point it is well to call attention to the fact that special tube selection is extremely undesirable. Directives from the headquarters of both the Bureau of Ships and the Signal Corps state that special selection of tubes will not be permitted but that equipments must meet performance requirements with any and all tubes which pass the JAN tests for the types involved. When a tube requires replacement in the field any tube of the correct type which is in stock *must* work satisfactorily. If the tube being replaced was specially selected by the equipment manufacturer in order to enable his equipment to pass acceptance tests, the tubes in Service stocks may not give good results and thus the equipment may be

inoperable at a time when it is most needed. The idea of including selected spare tubes with the equipment is not feasible due to the fact that tubes may have a shelf life and that it is very probable that the spares will become separated from the equipment during field operations. Special stocks of selected tubes are not wanted due to the additional handling and space required. Furthermore special selection is undesirable because of its adverse effect on production efficiency. It is definitely a military liability.

Fortunately the majority of radio engineers recognize the evils of special tube selection and are doing all possible to eliminate it by proper equipment and circuit design, but all design engineers are requested to give this their continued attention. The preceding remarks to the effect that care should be taken to insure that the JAN-1A Specification controls all important tube characteristics are, therefore, particularly applicable to the prevention of special tube selection.

The JAN-1A Specification provides for type approval of each tube type produced by each tube manufacturer. The purpose of requiring that a tube manufacturer secure type approval for each type that he wishes to sell to the Government is to provide a check on his ability to produce those tubes satisfactorily, and to allow the Services to check the accuracy and capabilities of his test equipment as well as the suitability of the tube for use by the Armed Services. Then, when the Contracting Officer places a contract with a manufacturer holding type approval for the required types, he knows that the contractor can actually make and test those types satisfactorily.

Complete information regarding the obtaining of tube type approval is contained in Form TAI-1, which may be secured from either the Bureau of Ships or the Signal Corps Standards Agency. However, a brief description of the procedure will be given.

1. The manufacturer writes to the Signal Corps Standards Agency in triplicate stating that he wishes to submit type JAN-______ for type approval. This letter must include complete information concerning the tube, test data, photographs, and other data required by Form TAI-1.

2. The Signal Corps Standards Agency advises him of the laboratory which will perform the tests and authorizes him to ship the tubes there.

3. The laboratory performs the tests and submits complete data with recommendations to the Signal Corps Standards Agency. The laboratory may correspond with the manufacturer, if necessary, during the performance of the tests.

4. The Signal Corps Standards Agency notifies the manufacturer of the outcome of the tests and issues type approval certificate, if warranted. If not warranted, additional samples may be submitted after the manufacturer has provided proof that suitable changes have been made in the tubes.

It should be understood that each manufacturer should obtain type approval for each type of his own manufacture that he wishes to sell for ultimate use by either the Signal Corps or the Navy. However, a manufacturer cannot secure approval for a type which he purchases from another manufacturer and resells. The JAN-1A Specification describes the correct method for marking tubes to indicate that their manufacturer holds type approval for them. It may be well to call attention to a basic requirement that, regardless of who sells the tube, the type-approval marking must be that of the actual manufacturer of the tube and is to be used only when that manufacturer has been granted type approval for that type.

An important function of the JAN-1 Tube Subcommittee is the handling of requests for waivers to the JAN-1A Specification. A tube manufacturer may find that he cannot make a required test due to lack of equipment; an equipment contractor may encounter a long delay in securing a particular type if all JAN tests are complied with, but can secure quick delivery of that type if it is tested to broader limits which still permit it to give proper operation in the particular equipment concerned. Such situations often arise. If they are referred to the Bureau of Ships or the Signal Corps Standards Agency through proper channels, the JAN-1 Tube Subcommittee will consider them and decide whether the tube will meet the requirements of the JAN-1A Specification. Decisions required to keep up production may be secured within a few hours by telephone and teletype concurrence of the JAN-1 Tube Subcommittee members.

Two types of waivers known as Type 1 and Type 2 are provided. Type 1 waiver may apply to any or all tube manufacturers who are concerned with the tube type affected and allows the use of the JAN marking. The Type 2 waiver applies to a particular manufacturer, a particular equipment, a particular contract, a specified number of tubes, and/or a specified length of time. Tubes supplied under a Type 2 waiver cannot be marked JAN.

Requests for Type 1 waivers should be sent directly to the Signal Corps Standards Agency or Bureau of Ships, Code 930A. Requests for Type 2 waivers should be made through the channels normally used in requesting approval for technical changes in contracts.

Over a year ago, the Army and Navy jointly issued a Preferred List of Vacuum Tubes. The first revision was made early in 1943 (the current revision is dated February 15, 1944) and included both a confidential and an unclassified listing. The object of these Preferred Lists, which must be aimed for in all developments, is to reduce number of types used by the Services. With fewer types required there are important savings to the Services in space and manpower for handling stock; also tube manufacturers can, by concentrating on the production of those types, achieve greater production efficiency, higher quality, and lower shrinkage. The JAN-1 Tube Subcommittee members revise these Preferred Lists from time to time on the basis of each type's wideness of application, ease of production, and available production

capacity. If a development requires tube characteristics which can be shown to be lacking in any of the preferred types, a waiver of the Preferred Lists may be requested from either the Bureau of Ships or Signal Corps Standards Agency. Thus, the fact that the use of the Army-Navy Preferred List of Vacuum Tubes is mandatory does not serve as an obstacle to new developments in tubes and equipments.

Since some of the tubes now in use and being developed incorporate features, the knowledge of which might benefit the enemy, it is necessary to assign a security classification to them. The JAN-1 Tube Subcommittee members are, by virtue of their close contact with personnel concerned with tube work for the government, fully cognizant of the tube features which should be withheld, so are called upon to recommend types for classification, or for removal from their security classification, after there is no longer need for restricting information concerning them.

Although the RMA continues to assign type numbers for new tubes, the JAN-1 Tube Subcommittee members work in co-operation with RMA regarding arrangements and plans for handling this work.

The most recent addition to the work of the subcommittee is that concerned with the use of the JAN-1A Specification by Canada. The Canadian Services require the use of this specification in all contracts using tubes which are placed after January 1, 1944. The Canadian Services will administer the specification through a joint Industry-Government committee and will co-ordinate all actions with the JAN-1 Tube Subcommittee in order to assure a yet broader standardization of tubes.

Thus has been achieved a single specification for use with all tubes procured by the United States Signal Corps, the United States Navy, and the Canadian Armed Services. After the transition stage, during which existing tube stocks will be used up, all tubes purchased and stocked by the organizations just named will be fully interchangeable. This one factor alone is of tremendous importance in the field, where replacements are needed in a hurry. Army tubes will now work in Navy equipments and vice versa, without resulting in loss of valuable time while maintenance personnel try to find out if, for example, the type VT-100 called for in a certain Army set can be replaced by Navy type 38807. Now, regardless of which organization purchased the tubes, the type number JAN-807 has taken the place of those two former designations, and pooling of Army and Navy stocks causes no maintenance problems. Of course full use of JAN type numbers is being made in marking equipment sockets.

Furthermore, by pooling their requirements in the JAN-1A Specification the Army and Navy have been able to improve the quality of some of their tubes. The manufacturer no longer is obliged to make the same tube meet two different sets of limits, so he can now concentrate on making that tube better. Already reports have been received from fighting fronts telling of the superior characteristics and quality of American tubes. The value of this joint effort is becoming manifest.

Through the use of preferred types, the tube manufacturer is further aided by the opportunity to specialize on a few types. Due to the existence of older equipments which required a wide variety of types, the full effect of the Preferred-List program cannot be felt for some time, but there are few who do not subscribe to the program as one of real value.

Lastly, the Services are working more closely with tube manufacturers than ever before. The manufacturers are consulted in all tube matters which affect tube production and it seems safe to say that there is greater co-operation in this respect than ever before. This is as it should be, and the JAN-1 Tube Subcommittee members are happy to have had the opportunity to aid in bringing this about. Their aim is to give the United States Army and Navy the world's best tubes in the vast quantities required. We appreciate the fine co-operation of industry in helping us to achieve this. If, in addition to meeting the requirements of this war, this specification is so satisfactory that it will be kept as standard in the commercial production which follows the victory, we shall feel that we have been successful. The Services are desirous of co-operating with industry to continue this tube standardization after the war.

In closing, it is interesting to quote from a report on Lessons in Signal Operation from Burma. "American tubes are superior because they are (1) sturdier and (2) standardized and interchangeable."

# The Institute of Radio Engineers
INCORPORATED

## SECTION MEETINGS

| | | | | |
|---|---|---|---|---|
| ATLANTA | CHICAGO | CLEVELAND | DETROIT | LOS ANGELES |
| September 15 | September 15 | September 28 | September 15 | September 19 |
| NEW YORK | PHILADELPHIA | PITTSBURGH | PORTLAND | WASHINGTON |
| September 6 | October 5 | October 9 | August 14 | September 11 |

## SECTIONS

**ATLANTA**—Chairman, Walter Van Nostrand; Secretary, Ivan Miles, 554—14 St., N. W., Atlanta, Ga.

**BALTIMORE**—Chairman, G. J. Gross; Secretary, A. D. Williams, Bendix Radio Corp., E. Joppa Rd., Towson, Md.

**BOSTON**—Chairman, R. F. Field; Secretary, Corwin Crosby, 16 Chauncy St., Cambridge, Mass.

**BUENOS AIRES**—Chairman, L. C. Simpson; Secretary, I. C. Grant, Venezuela 613, Buenos Aires, Argentina.

**BUFFALO-NIAGARA**—Chairman, Leroy Fiedler; Secretary, H. G. Korts, 51 Kinsey Ave., Kenmore, N. Y.

**CHICAGO**—Chairman, A. B. Bronwell; Secretary, W. O. Swinyard, Hazeltine Electronics Corp., 325 W. Huron St., Chicago, Ill.

**CINCINNATI**—Chairman, J. L. Hollis; Secretary, R. S. Butts, Box 1403, Cincinnati 2, Ohio.

**CLEVELAND**—Chairman, A. S. Nace; Secretary, Lester L. Stoffel, 1095 Kenneth Dr., Lakewood, Ohio

**CONNECTICUT VALLEY**—Chairman, R. E. Moe; Secretary, L. A. Reilly, 989 Roosevelt Ave.; Springfield, Mass.

**DALLAS-FORT WORTH**—Chairman, D. J. Tucker; Secretary, P. C. Barnes, WFAA-WBAP, Grapevine, Texas.

**DAYTON**— Chairman, I. H. Gerks; Secretary, Joseph General, 1319 Superior Ave., Dayton, 7, Ohio.

**DETROIT**—Chairman, R. A. Powers; Secretary, R. R. Barnes, 1411 Harvard Ave., Berkley, Mich.

**EMPORIUM**—Chairman, H. D. Johnson; Secretary, A. Dolnick, Sylvania Electric Products, Inc., Emporium, Pa.

**INDIANAPOLIS**—Chairman, A. N. Curtiss; Secretary, E. E. Alden, WIRE, Indianapolis, Ind.

**KANSAS CITY**—Chairman, A. P. Stuhrman; Secretary, R. N. White, 4800 Jefferson St., Kansas City, Mo.

**LOS ANGELES**—Chairman, L. W. Howard; Secretary, Frederick Ireland, 1000 N. Seward St., Hollywood, 38, Calif.

**MONTREAL**—Chairman, F. S. Howes; Secretary, J. A. Campbell, Northern Electric Co., 1261 Shearer St. Montreal, Que., Canada.

**NEW YORK**—Chairman, Lloyd Espenschied; Secretary, J. E. Shepherd, 111 Courtenay Rd., Hempstead, L. I., N. Y.

**PHILADELPHIA**—Chairman, W. P. West; Secretary, S. Gubin, RCA Victor Division, Radio Corporation of America Bldg. 8–10, Camden, N. J.

**PITTSBURGH**—Chairman, B. R. Teare; Secretary, R. K. Crooks, Box, 2038, Pittsburgh, 30, Pa.

**PORTLAND**—Chairman, W. A. Cutting; Secretary, W. E. Richardson, 5960 S.W. Brugger, Portland, Ore.

**ROCHESTER**—Chairman, O. L. Angevine, Jr.; Secretary, G. R. Town, Stromberg-Carlson Co., Rochester, N. Y.

**ST. LOUIS**—Chairman, N. J. Zehr; Acting Secretary, C. H. Meyer, KFUO, 801 DeMun Ave., St. Louis, Mo.

**SAN FRANCISCO**—Chairman, W. G. Wagener; Secretary, R. V. Howard, 225 Mallorca Way, San Francisco, Calif.

**SEATTLE**—Chairman, F. B. Mossman; Secretary, E. H. Smith, Apt. K, 1620—14 Ave., Seattle, 22, Wash.

**TORONTO**—Chairman, R. G. Anthes; Secretary, J. T. Pfeiffer, Erie Resistor of Canada, Ltd., 128 Peter St., Toronto, Ont., Canada.

**TWIN CITIES**—Chairman, E. S. Heiser; Secretary, B. R. Hilker, KSTP, St. Paul Hotel, St. Paul, Minn.

**WASHINGTON**—Chairman, J. D. Wallace; Secretary, F. W. Albertson, c/o Dow and Lohnes, E St. N. W., between 13th and 14th Sts., Washington, D. C.

# Institute News and Radio Notes

## Constitutional Amendment Section

PROPOSED INCREASE IN DUES

In the June issue of the PROCEEDINGS, a brief outline was given of the reasons for proposing an increase in dues. It was promised that more information would be supplied in later issues.

FINANCIAL DIFFICULTIES

In 1930 the Institute membership was 6600. In 1934 it had dropped to 4400. During that period the Institute operated at a deficit every year with the Board making every effort to keep expenses down. Since the Board could avoid a deficit only by retrenching, economies were instituted. In 1934 the PROCEEDINGS contained only 64 per cent of the number of pages appearing in 1930. This retrenchment in expenditure on the PROCEEDINGS resulted in balancing the budget in 1935, but brought about its own difficulties. The Board then began increasing the number of pages in the PROCEEDINGS to shorten publication delay, and did not replace members who left the office staff as a balancing economy. This brought about other troubles. By the latter part of 1941 it was obvious the Institute was in financial difficulties.

The Board then authorized an advertising campaign. This, and only this, has avoided serious difficulties for the Institute. Advertising is now relatively easy to get, thanks to large government expenditures for radio, and to certain other circumstances. Our advertising is now largely limited by our ration of paper. This comparatively easy-to-get advertising, however, is an insecure foundation for plans. It may disappear with the termination of the war. It cannot be depended upon.

INCREASED COSTS

The financial difficulties just enumerated were based upon prewar wage levels. It will be news to no one that wage levels have risen since 1941. The Institute has had to meet these new wage levels. All indications are that wage levels will not fall after the war. Other costs, such as paper, printing, and rent, have also increased. The prewar difficulties, the new wage levels, and other costs must all be met if the Institute is to function after the war as well as before. The advertising income is uncertain. Only by an increase in dues can the members be assured of the continued functioning of the Institute on the prewar level.

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17 May, 1944

*The Editor*
*Proceedings of the IRE*
*New York, New York*

The letter from the Chairman of the Montreal Section, in the May PROCEEDINGS, proposing a change from the recently adopted membership grade names to certain other names, arouses memories. When the Board and special committees were considering the matter, during a period of over a year, the arguments presented were gone over time and again.

I am glad that the Montreal Section accepts all features of the change except that of names. The names are really pretty much a matter of personal opinion, but I should like to point out that the radio engineering profession has a different problem from those of all other engineering professions. We have thousands of people interested in radio and wanting to join the Institute who are not in the profession. Oppositely, I maintain that any man who has been in the profession long enough to qualify as a "Member" (present designation) is entitled to being called a member of IRE, and I do not think it is lowering the prestige of the Institute so to call him. On the other hand, it lowers the prestige due him to be called some uncertain kind of a member, qualified by some term like Associate or Affiliate whose meaning is not exact and may be anything—higher or lower than plain member, except by arbitrary definition.

So I submit that any one actually in the profession "for keeps" is entitled to be called a Member, without any qualification. When he gets high enough, he can be a Fellow. Before that he should be something which is clearly higher than Member, and the best term which came out of much thought on the subject was Senior Member.

Incidentally, it is interesting to note what the dictionary definitions are for the terms Associate and Affiliate:

> Associate is "a companion, partner, mate, fellow, confederate, accomplice, ally."
>
> Affiliate is "one adopted, received into a society or family as a son, with established paternity, as of a bastard child."

If we are to worry more about names, let's be exact, and choose words which name exactly what we mean—which is already accomplished by what we have, and is not accomplished by indefinite terms like Affiliate.

I recommend strongly, if this question again comes to ballot, that each member take the trouble to read all the communications on the subject which appeared in the PROCEEDINGS for May and June, 1943.

A. F. VAN DYCK
Lieutenant Commander
United States Naval Reserve

# Board of Directors

***May 3 Meeting:*** At the regular meeting of the Board of Directors on May 3, 1944, the following were present: H. M. Turner, president; R. A. Hackbusch, vice-president; S. L. Bailey, I. S. Coggeshall, Alfred N. Goldsmith, editor; R. F. Guy, R. A. Heising, treasurer; C. B. Jolliffe, F. B. Llewellyn, Haraden Pratt, secretary; H. J. Reich, B. J. Thompson, L. P. Wheeler, W. C. White, and W. B. Cowilich, assistant secretary.

***Executive Committee Actions:*** The actions of the Executive Committee meeting on April 4, 1944, which had been described in minutes mailed to the Board of Directors and subsequently amended, were ratified.

***Membership:*** The following applications for membership, which were approved by the executive Committee, were accepted: for transfer to Senior Member grade, G. P. Adair, O. M. Dunning, R. K. Honaman, Cullen Moore, A. C. Peterson, Jr., H. D. Seielstad, and D. B. Smith; for admission to Senior Member grade, L. N. Brillouin, W. T. Cooke, H. F. Fruth, G. H. Huber, Walther Richter, M. K. Taylor, and J. F. Willenbecher; for transfer to Member grade, R. C. Bailey, G. V. Brunner, F. L. Burroughs, C. H. Campbell, William Dickinson, J. E. Doane, C. B. Eckel, B. S. Ellefson, E. M. Goldberger, Harry Halinton, B. J. Kingsburg, P. B. Laeser, Nathan Marchand, E. E. Overmeir, C. B. Plummer, A. L. Schoen, E. W. Vaughn, W. A. Weiss, and Halley Wolfe; for admission to Member grade, E. W. Allen, Jr., N. G. Anton, T. J. Carroll, C. I. Cronburg, Jr., E. J. Daubaras, C. R. Evans, C. F. Kocher, R. W. Liska, and S. J. McDermott; Associate grade, 152; and Student grade, 76.

***Committees:*** The following members, as recommended by the Executive Committee, were appointed to these committees:

CIRCUITS

| | |
|---|---|
| C. Brunetti | J. B. Russell, Jr. |
| L. A. Kelley | W. N. Tuttle |
| C. Nietzert | H. A. Wheeler |

PAPERS

E. W. Herold

PAPERS PROCUREMENT

D. D. Israel, *General Chairman*
W. L. Everitt, *Vice-chairman*

GROUP CHAIRMEN

| | |
|---|---|
| J. E. Brown | J. K. Johnson |
| E. J. Content | C. J. Madsen |
| Harry Diamond | D. H. Moore |
| E. T. Dickey | H. A. Wheeler |

W. C. White

***Sections:*** Treasurer Heising, as chairman of the Sections Committee, reported that with the additional information recently received, the petition for the formation of a Williamsport Section now carries the signatures of 29 members who are of the required grades, and that all such members are in good standing and have their addresses in Lycoming and Clinton Counties, the Section territory requested.

## National Electronics Conference

*ELECTRONICS CONFERENCE PLANNED AS NATIONAL FORUM*

Scheduled to be held at the Medinah Club, 505 N. Michigan Ave., Chicago, Illinois, on October 5, 6, and 7, the National Electronics Conference will offer a symposium of papers covering the communications, industrial, measurements, and medical applications of electronics. Every attempt is being made, within the limits of national security, to assure that the papers to be presented will not only be outstanding in their field, but will also reflect recent progress. Nationally known speakers have been invited to address the Conference at its banquet on October 5, and at the luncheons on October 5 and 6.

Described as "a national forum for electronic developments and their engineering application" by Dr. J. E. Hobson, director of the School of Engineering of the Illinois Institute of Technology, and chairman of the Executive Committee, the Conference has as its aim, a means of providing free discussions of the applications of electronics in various fields of technical endeavor.

The program for the Conference is under the direction of Professor A. B. Bronwell, chairman of the Program Committee. While the general plan of the program has been outlined, those desiring to submit papers at the Conference are invited to communicate with Professor Bronwell, Technological Institute, Northwestern University, Evanston, Ill.

Advance registration may be made with Professor P. G. Andres, Illinois Institute of Technology, 3300 Federal St., Chicago.

A well-rounded program of activities is assured by the Illinois Institute of Technology, Northwestern University, the Chicago Section of The Institute of Radio Engineers, and the Chicago Section of the American Institute of Electrical Engineers, sponsors of the National Electronics Conference.

With the Constitutional requirements thus met, and on the recommendation of the Executive Committee, the Board of Directors officially approved the establishment of the Williamsport Section.

***National Electronics Conference:*** The recommendation of the Executive Committee relative to the publication of a notice of the National Electronics Conference in Chicago in October, 1944, was approved.

***Constitution and Bylaws:*** The rewording of the following Articles were discussed and in most cases unanimously approved by the Board: Article I, Section 2; Article II, Section 1A; Article IV, Section 1; Article VII, Section 1; Article IX, Sections 1 and 2; and Article X, Section 2.

Proposed Bylaws Amendments:

Sections 37 and 40: It was unanimously adopted to replace the word "residing" with the words "with mailing addresses" in Sections 37 and 40.

Section 45: The revised amendment of this Bylaws Section was adopted as follows:

"Sec. 45—The standing committees, each of which shall normally consist of five or more persons, shall include the following: Admissions, Appointments, Awards, Board of Editors, Constitution and Laws, Education, Executive Committee of the Board of Directors, Investments, Membership, Nominations, Papers, Public Relations, Sections, Tellers, Annual Review, Antennas, Circuits, Electroacoustics, Electronics, Facsimile, Frequency Moduation, Radio Receivers, Radio Transmitters, Radio Wave Propagation, Standards, Symbols, Television.

"These committees shall be advisory to the Board of Directors on those matters which are reasonably described by the committee names, except as defined in these Bylaws.

"The terms of appointments of the Admissions, Awards, Board of Editors, Constitution and Laws, Education, Investments, Membership, Nominations, Papers, Public Relations, Sections, and Tellers Committees shall start with the first day of the month following appointment and continue until the date the succeeding terms of appointments take effect. The Board may specifically advance or delay the terminating date of any committee and the starting date of a succeeding committee.

"The Board shall make appointments to the following committees: Annual Review, Antennas, Circuits, Electroacoustics, Electronics, Facsimile, Frequency Modulation, Radio Receivers, Radio Transmitters, Radio Wave Propagation, Standards, Symbols, and Television, each year between January first and May first and the terms of appointments shall be from May first of the year when the appointments are made until April thirtieth of the following year. Additional appointments may be made to fill vacancies or to care for special cases as the need arises, with the term of the appointment expiring April thirtieth."

***Canadian Membership:*** Vice-President Hackbusch reported on the meetings with the Canadian Wartime Labour Relations Board, held on April 11 and 12, 1944, at Ottawa, at which various technical and professional groups were represented, and which he attended as the representative of the Canadian membership of the Institute.

It was stated by Mr. Hackbusch that the Canadian situation is favorable to the organization of separate collective-bargaining groups for professional engineers.

***T. L. Eckersley:*** On the recommendation of the Executive Committee, the Fellow award was granted to Mr. Eckersley of England.

# Executive Committee

***May 2 Meeting:*** The Executive Committee meeting, held on May 2, 1944, was attended by H. M. Turner, president; E. F. Carter, Alfred N. Goldsmith, editor; R. A. Heising, treasurer; Haraden Pratt, secretary; and W. B. Cowilich, assistant secretary.

### Meetings

*Winter Technical Meeting—1945:* Dr. Austin Bailey has accepted the chairmanship of the 1945 Winter Technical Meeting Committee.

*American Institute of Electrical Engineers:* The A.I.E.E. plans to hold its Winter Technical Meeting on January 22 to 26, 1945; and consideration was given to co-ordinating this meeting with that of the I.R.E.

### Standards

*Definitions of Guided Waves:* Unanimous approval was given to the Standards Committee recommendation that this report, as submitted by the Technical Committee on Radio Wave Propagation, be printed as a separate publication and in a quantity sufficient to provide copies to all members and subscribers and a surplus stock.

*Piezoelectric Crystals:* It was agreed to publish this report as a separate publication, with any further editorial changes that may be made by Professor Cady of the Piezoelectric-Crystals committee, and to provide copies to all members and subscribers and a surplus stock, as recommended by the Standards Committee.

## Illinois Radio Engineers

All radio engineers residing in the state of Illinois or practicing engineering in that state should apply for licenses on or before July 31, 1944, in order that they may obtain licenses without examination in case the Supreme Court upholds the contested Illinois Professional Engineering Act. Application blanks may be secured from the secretaries of various engineering organizations; Mr. Frank F. Fowle, president of the Illinois Engineering Council, 35 E. Wacker Dr., Chicago, Ill.; or Mr. A. W. Graf, Secretary, Chicago Section, I.R.E., 135 S. LaSalle St., Chicago, Ill.

Failure to file applications by July 31, 1944, will mean that the applicant will have to pass a written or oral examination in engineering to obtain a professional engineering license at some subsequent date if the Supreme Court validates the contested Act. The fee in the latter case is $20.00 instead of $10.00 required without examination.

# Inventive Problems of Military Interest

The following list of new problems, the solutions of which would be helpful to the Military Services, is submitted to the readers of the PROCEEDINGS OF THE I.R.E. It constitutes a selection from a more comprehensive list, the selection being based on the likely interests of the readers of the PROCEEDINGS.

Ideas directed to the solutions of these problems should be addressed to The National Inventors Council, Department of Commerce, Washington 25, D. C. The Institute is advised that these communications will be promptly analyzed and that those which appear novel and promising will be placed in the hands of appropriate Military personnel.

*The Editor*

1. A relatively simple gage to measure the impulse of explosion blast, positive and negative phases to be determined separately but concurrently. It would be desirable if the duration of each phase could be determined in some simple manner.
2. Device to maintain or indicate, within 5 feet, the relation of an aerial camera to the vertical.
3. Reduction of glare from glass surfaces by *durable* coatings suitable for field application.
4. Optical method for determining the difference between an artificial green and a natural green.
5. Destruction and removal of obstacles to landing operations. Obstacles may be visible or concealed and may be off or on shore.
6. Location and destruction of concealed enemy emplacements, pillboxes, and similar strong points.
7. Methods of protecting our vehicles from the effects of enemy land mines.
8. Ingenious and simple decoy devices for purpose of confusing and misleading enemy.
9. Technical data as to strategic enemy targets such as chemical plants, explosive plants, power plants, etc.
10. A voice-transmitting gas mask which would permit the wearer's voice to be heard with clarity.
11. Methods of generating stable artificial fogs, and methods of dispersing artificial and natural fogs.
12. Protection against flame throwers.
13. *Noiseless* hand generator combined with a lightweight flashlight. The generator should be pumped at a rate of 40 revolutions per minute, and the light should be continuously brilliant and start on the first pump.
14. Plumb-bob type of generator to operate a light 6-volt radio. Generator probably operated by a reel winding up and letting out the plumb-bob.

# Books

## Radio Audience Measurement, by Matthew N. Chappell and C. E. Hooper

Published by Stephen Daye, Inc., 48 E. 43 St., New York 17, N. Y. 239 pages+5-page index. 43 illustrations. $8\frac{1}{2}\times5\frac{3}{4}$ inches. Price $3.50.

Broadcasting has become well established as a major method of instantaneous mass communication. It has the limitation of speaking in a single direction, that is, outward from the transmitting station to the multitude of broadcast listeners' homes. Thus the audience is not only invisible but also inaudible. Yet the sponsors of programs, the networks, and the individual stations over which the programs are radiated have a deep and direct interest in the reactions of the voiceless audience. Accordingly there has been devised a considerable group of diverse methods having as their aim the tracking down of the elusive audience reaction. The problem is not a simple one.

In the book, "Radio Audience Measurements," the main methods of questioning the audience are described in acceptable detail. Among those analyzed are the coincidental method (which involves an audience cross-sectional poll by telephone calls during the program period under study), the recall methods (which are based on inquiries made after the program period), and panel and fixed-sample methods (sometimes based on the use of mechanical recorders of the particular station to which the individual listeners are tuned at all times). The advantages and disadvantages of the various methods are presented.

The book is primarily of interest to networks, stations, sponsors, advertising agencies, and students of public-opinion trends. In this last connection, the book presents an incomplete but generally informative discussion of the statistical bases of public-opinion polling.

Engineers may be particularly interested in certain aspects of the book. One of these is the interrelation between the physical coverage of a broadcast station (as largely determined by the signal-intensity distribution in the area surrounding the station) and its audience coverage (which involves many other and sometimes evasive factors). The mechanical type of recorder of the listener's station selections serves as a partial link between these methods, and presents the possibility of further engineering development.

Recently there has been brought to the attention of an appropriate Panel (Panel 13) of the Radio Technical Planning Board the matter of providing experimental facilities (principally frequency allocations) for a new service termed "centercasting." This service involves two-way communication between a central transmitting and receiving station, on the one hand, and, on the other hand, a considerable number of surrounding voter and respondent stations (each also having a transmitter and a receiver). Thus questions can be asked of the audience, and the answers of the listeners can be speedily

obtained by radio. The process is applicable not only to broadcasting itself, as is highly appropriate, but also to many commercial, political, and sociological inquiries. Possibly centercasting will find its place in later editions of this book.

Written in a clear and instructive style, the book in its present form presents a reasonably comprehensive picture of the field. Although its authors might be expected to have a favorable bias toward the methods of a particular audience-measurement group, they appear to have preserved in the main a sufficiently judicial attitude in their treatment of audience polling.

Alfred N. Goldsmith
Consulting Engineer
New York, N. Y.

## Basic Radio, by C. L. Boltz

Published (1944) by The Ronald Press Company, 15 E. 26 St., New York 16, N. Y. 269 pages+3-page index+viii pages. 166 figures, $5\frac{1}{4}\times7\frac{3}{4}$ inches. Price, $2.25.

This book is offered by the publisher as an elementary textbook covering that basic knowledge without which one cannot go on in a serious study or advanced work in radio. It is, therefore, unfortunate that it defeats to some extent the above purpose by containing incorrect explanations of the theory. Furthermore, the author exhibits a certain air of disdain for some of the theoretical and mathematical aspects of electrical engineering that cannot help but militate against advanced study on the part of the reader. For example, on page 52, he states concerning Kirchhoff's laws, "We include these laws here because the textbooks so often quote them, and people who set examinations delight in asking for them. In practice we apply the rules we have already learned earlier direct to the circuit we are considering." Indeed, on page 54, he further states, "By using our knowledge of resistances in series and parallel we can calculate the equivalent resistance of any network," which is a neat trick if it can be applied directly, for example, to a bridge circuit without invoking the aid of the delta-wye transformation. Other errors occur, such as on page 219, the statement that the class B amplifier cannot be used to amplify a modulated carrier because detection takes place, or on page 242, the confusing of direct-current input to a tube with plate dissipation; and as a final example, on page 256, et al., a somewhat fantastic explanation of the action of a transmission line.

The book is simple enough but it is marred throughout by the excessive use of the word "so." A typical example of this will be found on page 217. It is clearly not intended for engineers and it would be better suited for the beginner if it were rid of its inaccurate explanations.

A further improvement would be to amplify the book somewhat to include material on superheterodyne receivers and possibly some material on ultra-high-frequency technique. However, the printing is well done and the diagrams are, in general, satisfactory.

In conclusion the reviewer cannot help but express the wish that elementary engineering texts were written by world-famous engineers who have matured to the point where they have penetrated the jungle of detail and once more see the broad aspects of the subject.

Albert Preisman
Capitol Radio Engineering Institute
Washington, D. C.

## Experiments in Electronics and Communication Engineering, by E. H. Schulz and L. T. Anderson

Published (1943) by Harper and Brothers, 49 E. 33 St., New York, N. Y. 377 pages +3-page index+xii pages. 300 figures. $6\frac{1}{2}\times9\frac{1}{4}$ inches. Price $3.00.

An integrated laboratory text for the fields of electronics, communication networks, radio, and ultra-high-frequency phenomena should be a distinctly useful type of book. "Experiments in Electronics and Communication Engineering" is such a text. Originally prepared for the Signal Corps Training Schools at the Illinois Institute of Technology, the experiments have been revised to a form usable in undergraduate college and Engineering Science Management War Training courses as well. In spite of the speed with which it must have been prepared, the book is neither careless nor incomplete. An occasional minor error occurs, mainly in the diagrams; but the lithoprinted form suggests that a second printing will be without most of these.

The four major fields are covered in 108 experiments, each complete in itself but allowing easy modification, expansion, or omission. Two initial chapters contain no experiments, but discuss laboratory technique, procedure, and data presentation and equipment, circuit components, and instruments. Following these are two introductory groups of experiments on direct-current and alternating-current circuits. The more advanced work begins with chapter 5, on network theory and four-teminal networks. Subsequent chapters are on electron tubes, cathode-ray tubes and circuits, power supplies, amplifiers, oscillators, modulation and detection, radio receivers, radio transmitters, and transmission lines, wave guides, and radiation. Each experiment contains a resume of the necessary theory and references to several standard texts. The list of apparatus and suggested procedure is flexible; unnecessarily detailed treatment is avoided in favor of the use of initiative and a practical approach by the student. There is a full quota of conventional experiments, and in addition a few unusual items like the study of fields in a triode with a water model.

Because of its flexibility, this text should prove useful in any one of the fields covered or in a program combining them all. The clarity and up-to-date coverage of almost the entire communication option recommend it for normal civilian needs as well as the war training programs.

O. L. Updike
University of Illinois
Urbana, Illinois

## Practical Radio and Electronics Course, Three Volumes, Prepared by M. N. Beitman

Published (1943) by Supreme Publications, 328 S. Jefferson Street, Chicago, Illinois.
Vol. I—171 pages+1-page index. 192 figures. $8\frac{1}{2}\times10\frac{3}{4}$ inches.
Vol. II—267 pages+1-page index. 109 figures. $8\frac{1}{2}\times10\frac{3}{4}$ inches.
Vol. III—99 pages+1-page index. 86 figures. $8\frac{1}{2}\times10\frac{3}{4}$ inches.
Price $3.95 for 3 vols.

These home-study course manuals have been prepared by careful analysis of the technical literature that describes commercial equipment used in the radio field. The compilation in these three sections is quite extensive and will probably serve best to help students who have only a classroom knowledge of radio theory, as to the details of commonplace apparatus he is likely to encounter. Editorial comments accompany each description as marginal notes. The subject matter seems well selected for this series.

Ralph R. Batcher
Radio Consultant
St. Albans, L. I., New York

## The Radio Amateur's Handbook

Published (1944) by The American Radio Relay League, West Hartford, Connecticut. 480 pages+175-page catalog section+10-page index. 1125 figures+125 charts and tables+175 basic formulas. $6\frac{1}{2}\times9\frac{1}{2}$ inches. Price, $1.00 in the United States; $1.50 elsewhere.

It seems almost futile to review a book which in twenty previous editions during a period of about eighteen years has enjoyed a distribution of over a million copies. Such is the A.R.R.L. Handbook, the scope of which is the broad field of amateur radio. Taking the abilities of its authors and editors for granted, the frequent revisions to keep its contents constantly up to date and its low price are the foremost reasons for the Handbook's continued pre-eminencne.

The neophyte, with only a desire to learn about high-frequency (112 to 224 megacycles with brief glimpses up to 750 megacycles) radio theory and amateur practice, needs no other text and can get along without a teacher, as long as he has a Handbook. Its straightforward and simple presentations of radio principles and practices stem from years of writing for the amateur under conditions where a poorly done job is exposed promptly by an avalanche of mailed questions which come uncomfortably close to the desk of the author.

The war has shifted the "balance" of the Handbook to put greater emphasis on theory. This has suited it better for the teaching of radio trainees in the Armed Forces, a not inconsequential role that it now plays.

In its constructional pages, a new entry is a chapter on carrier-current communications, a field still open to amateurs who are not now permitted to radiate power into space. The other activity providing operating privileges, the War Emergency Radio Service, has been treated in a completely rewritten chapter of almost fifty pages.

An outstanding text for either individual or group training, the A.R.R.L. Handbook continues to be the best dollar's worth of technical radio literature available to anyone.

Harold P. Westman
Institute of the Aeronautical Sciences
New York 20, N. Y.

## Fundamentals of Radio Communications, by Austin R. Frey

Published (1944) by Longmans, Green and Co., Inc., 55 Fifth Ave., New York, N. Y. 385 pages+7-page index+xii pages. 201 illustrations. $5\frac{3}{4} \times 9\frac{1}{4}$ inches. Price, $4.00.

This new book is one which should interest engineers and physicists who want to study basic radio circuits and their behavior, and also those who want to review the subject to improve their knowledge in this technical field. The author presumes that the reader knows mathematics through calculus and can use differential equations, hyperbolic functions, and Bessel functions with some facility. In addition to this mathematical background, the reader must have a good knowledge of the fundamental principles of electric circuits, such as a working knowledge of the use of vectors and the laws of current and voltage relationships in alternating-current circuits. A person with this background will find the book clearly written and the material well organized. Time and space are not wasted in the development of mathematical expressions; but they are freely used wherever they convey the essential information more clearly and succinctly than words.

The first chapter (42 pages) is devoted to the analysis, by the use of complex numbers, of resonant and coupled circuits as they are used in vacuum-tube amplifiers; while the second and third (60 pages) delineate the characteristics of vacuum tubes and introduce the association of the tubes with the basic circuits that are used in the several vacuum-tube stages of transmitters and receivers. The next two chapters (110 pages) explain the action of the standard types of alternating-frequency and radio-frequency amplifiers, using both analytical and graphical developments. The sixth chapter (46 pages) is devoted to oscillatory circuits, starting with the spark circuit because of its utility in explaining transients in communication circuits, through the conventional low-frequency generators to several ultra-high-frequency types. The classical solution of the differential equation for the series circuit of $L$, $C$, and $R$ is given to explain the spark circuit and the transient response of such a circuit in general. Chapter seven (66 pages) covers modulation and demodulation; presenting the usual trigonometric approach for amplitude modulation and Bessel function approach for frequency modulation. The eighth chapter (22 pages) is confined to a brief treatise on radio-frequency transmission lines and uses hyperbolic functions for the analytical evaluation of line characteristics. Chapter nine (34 pages) is entitled "Radiation" and deals primarily with radiation formulas and patterns. Problems are included at the end of each chapter by means of which the reader may test his understanding of the material he has read. However, the lack of solved examples, or answers to the problems given, limits their utility except for classroom use.

As may be surmised from the number of pages devoted to the chapters covering the various topics, the treatment of each is brief, as compared to some of the older books in this field written on the same mathematical level. Except for a block diagram of a transmitter on page 3 and one of a superheterodyne receiver on page 280, no drawings are given to show the organization of the several types of circuits into commercial sending and receiving apparatus. No illustrations or drawings of coils, condensers, or commercial equipment are included; no space is devoted to microphones or loudspeakers, and no attempt is made to cover any of the multitude of special circuits based on the long-known and basic circuits covered in this book. This paragraph is not intended as an adverse criticism, but since most books in this field endeavor to cover the topics listed above, their omission is worthy of mention.

The author states in his preface that his book is devoted to the fundamental principles of radio communications and limited to the more important types of circuits and circuit theory. As the chapter headings indicate, the book presents and analyzes the basic circuits that appear in transmitters and receivers in radio communication. Dr. Frey has covered this field admirably, writing concisely and clearly.

George F. Maedel
R.C.A. Institutes, Inc.
75 Varick St.
New York 13, N. Y.

## Illustrated Technical Dictionary. Edited by Maxim Newmark

Published (1944) by The Philosophical Library, 15 E. 40 St., New York 16, N. Y. 352 pages+vi pages. 90 figures. $6\frac{1}{4} \times 9\frac{1}{4}$ inches. Price, $5.00.

The plan of this dictionary is to supply a wide and representative selection of officially approved terms and definitions standardized by various United States engineering, technical, trade, and industrial organizations, as well as by government agencies. The terms included were determined on the basis of whether that subject is fundamental in all fields of engineering and has current importance in present-day technical activity. The selection of terms dealing with standards and measurements, and practical applications of mathematics, chemistry, and physics has been quite extensive. Terms dealing with aeronautics, electricity, radio, the chemistry of synthetics and plastics, automotive mechanics, and shipbuilding have been included. A certain number of tables of technical abbreviations, standards and measurements, mathematical, scientific, and technical symbols, conversion tables, shop data, and formulas, has also been included.

Ralph R. Batcher
Radio Consultant
St. Albans, L. I., New York

# Contributors

Maurice Artzt

Maurice Artzt (A'26) was born in Tyler, Texas, on November 30, 1904. He received the B.S. degree in electrical engineering from the University of Texas in 1925, and then joined the radio department of the General Electric Company at Schenectady. While there he attended Union College and received the M.S. degree in electrical engineering in 1929. He has been with the Radio Corporation of America since 1930, first at Camden, and later at the RCA Laboratories in Princeton. Since 1928 his work has been largely concerned with the development of circuits and apparatus for facsimile services. Mr. Artzt is a member of Sigma Xi and Tau Beta Pi.

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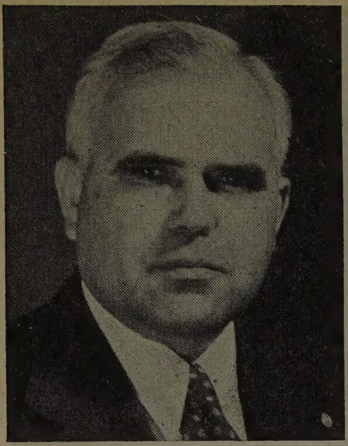
Edgar H. Felix



Harald Trap Friis

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Edgar H. Felix (J'17-A'19-SM'43) was born on March 29, 1898. He studied at New York University, Columbia University, and Sheffield Scientific School, Yale University. In 1918 Mr. Felix joined the Signal Corps as an engineer in the radio development section. In 1919 he became associate editor for *Aerial Age Weekly*. From 1922 to 1924 he was director of public relations for WEAF and the American Telephone and Telegraph Company. In 1924 he became technical director of the radio department of N. W. Ayer and Son, and from 1926 to 1941 he was radio consultant to broadcast stations, networks, and publishers. He served as radio director of the National Electrical Manufacturing Company from 1927 to 1929; contributing editor of *Radio Broadcast* 1926 to 1930; *Aero Digest*, 1930 to 1931; and *Advertising and Selling*, 1932 to 1935. In 1935 he became director of Radio Coverage Reports, which post he still occupies. He is the author of "Using Radio in Sales Promotion," published in 1937, and "Television, Its Methods and Uses," published in 1941. In 1942 he became a Captain in the Signal Corps, as assistant to the Chief Signal Officer, Fighter Control Section, and in 1943, with the rank of Major, he was assigned to the Radio Navigation Section.

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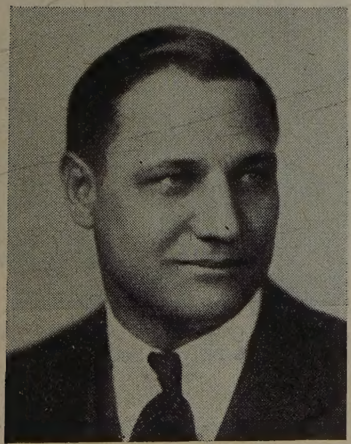

Walter Greer

Harald Trap Friis (A '18-M '26-F '29) was born in Denmark on February 22, 1893. He received the Electrical Engineering degree from the Royal Technical College in Copenhagen in 1916 and the Ph.D. degree in 1938. During 1916 he served as an assistant to Professor P. D. Pedersen, and in 1917–1918, he was a technical advisor at the Royal Gun Factory in Copenhagen. In 1919, Dr. Friis was made a Fellow of the American Scandinavian Foundation and did graduate work at Columbia University. He joined the research engineering staff of the Western Electric Company in 1920; this became the Bell Telephone Laboratories in 1925 and his work continued in that organization. In 1939 Dr. Friis received the Morris Liebmann Memorial Prize for his investigations in radio transmission including the development of methods of measuring signals and noise and the creation of a receiving system for mitigating selective fading and interference. He served as a director of the Institute from 1941 to 1944.

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J. Walter Greer (A'41), was born on January 21, 1906, at Adrian, Ohio. After receiving his B.S. degree in electrical engineering from Notre Dame University in 1929, he joined the Bell Telephone Laboratories. Studies on communication equipment and electronics were continued at the Laboratories and at Columbia University. Mr. Greer participated in research on electrical shock and in a program conducted jointly by the Bell Telephone Laboratories and Columbia Medical School. In 1935 he joined the RCA Manufacturing Company, working on the design and application of electron tubes. From 1936 to 1940 Mr. Greer was connected with the Tung-Sol Lamp Works as an engineer on the testing and processing of electron tubes. He joined the radio engineering section of the Civil Aeronautics Authority in 1940, and later in the same year transferred to the design branch, Bureau of Ships, Navy Department where he is in charge of the vacuum-tube section.

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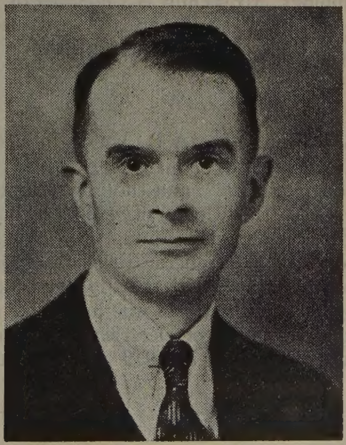
E. L. Hall (A'28) was born at Mansfield, Ohio, on April 20, 1893. He received the B.E.E. degree from the Ohio State University in 1918 and the E.E. degree from the same institution in 1929. He has been a member of the radio section, National Bureau of Standards since 1919, engaged in work on radio standards, measurements, and testing. He is a member of Sigma Xi.



E. L. Hall

Mack C. Jones

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Mack C. Jones (A'42) was born in Herrin, Illinois, on April 18, 1913. He received a B.S. degree in electrical engineering in 1935 from the University of Illinois. He was employed by RCA Manufacturing Company in 1935 and worked in the transmitter test department. In 1936 he transferred to the broadcast receiver engineering department and worked on the design of broadcast television and frequency-modulation receivers.

Since 1941 Mr. Jones has been in the communications and industrial engineering department associated with the design of Navy equipment.

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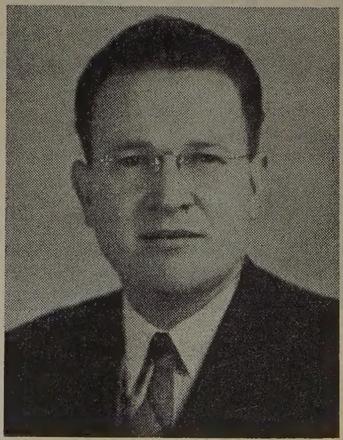
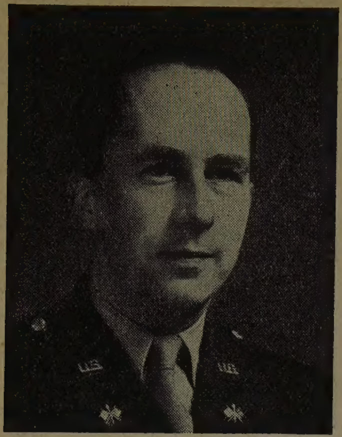
Vincent R. Learned, (S'38-A'40) was born in San Jose, California, January 21, 1917. In 1938 he received the degree of Bachelor of Science in Electrical Engineering from the University of California and in 1943 the Ph.D. degree from Stanford University. From 1938 through 1940, Dr. Learned was



Vincent Learned



C. W. Martel

with the engineering department of the McClatchy Broadcasting Company, Sacramento, California; from 1941 through 1942 he was a teaching and research assistant at Stanford University; and since January, 1943, he has been connected with the research laboratories of the Sperry Gyroscope Company, Inc., Garden City, New York, in the capacity of project engineer. He is a member of Tau Beta Pi, Eta Kappa Nu, and the Society of Sigma Xi.

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C. W. Martel was born on January 14, 1910, at Wells River, Vermont. He received the B.S. degree in physics from the Massachusetts Institute of Technology in 1931. During the following year he served as a laboratory assistant in the M.I.T. physics department. From 1936 until September, 1942, when he was called to active duty in the Signal Corps, he was connected with the Raytheon Production Corporation, where, for the latter two years he was assistant

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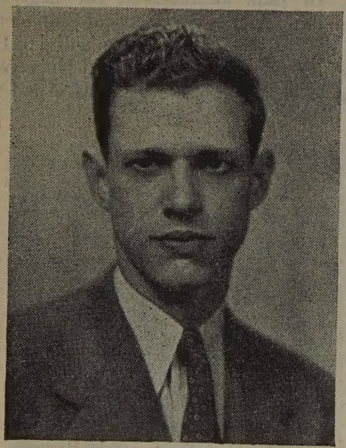
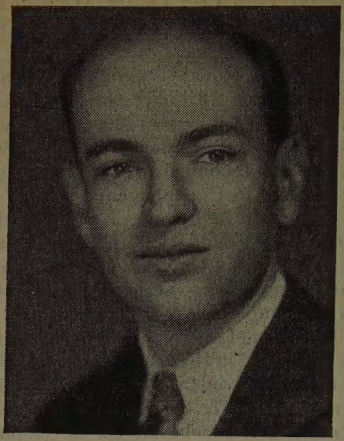


Edward Ralph Schatz



Elmer H. Schulz

manager of their small-tube division. Since entering the Service he has been assigned to the Aircraft Radio Laboratory, which he represents in electron-tube matters regarding Army Air Force radio equipment.

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Edward Ralph Schatz (A'44) was born at St. Mary's, Pennsylvania, on November 28, 1921. He received the B.S. degree in electrical engineering in 1942 from the Carnegie Institute of Technology. Continuing there he was appointed a Buhl Research Fellow in electrical engineering and obtained a Master's degree the following year. At present Mr. Schatz is an instructor in the electrical engineering department of the Carnegie Institute of Technology. He is a member of Eta Kappa Nu, Tau Beta Pi, and an Associate member of Sigma Xi.

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Elmer H. Schulz (A'38) was born at Lockhart, Texas, on October 30, 1913. He received the B.S. degree in electrical engineering in 1935 and the M.S. degree in electrical engineering in 1936 from the University of Texas. He was on the teaching staff of the electrical engineering department of the University of Texas from 1936 to 1942 when he joined the staff of the electrical engineering department of Illinois Institute of Technology. Mr. Schulz is a member of Tau Beta Pi and the American Institute of Electrical Engineers.

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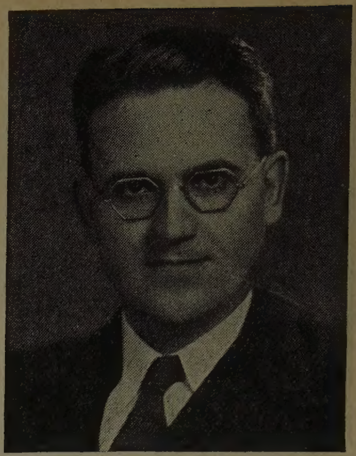
B. R. Teare, Jr. (A'41) was born in 1907 in Menomonie, Wisconsin. He received the B.S. degree in 1927, and the M.S. degree in 1928 from the University of Wisconsin. After another year of graduate work he was employed by the General Electric Company and was associated with the advanced course in engineering there. He was instructor and assistant professor of electrical engineering at Yale from 1933 to 1939 and completed



B. R. Teare, Jr.

the work for the degree of Doctor of Engineering in 1937. Dr. Teare now is Buhl Professor of electrical engineering at the Carnegie Institute of Technology. He is a member of Eta Kappa Nu, Tau Beta Pi, Phi Kappa Phi, and Sigma Xi, Fellow, A.I.E.E., and at present is chairman of the Pittsburgh Section of the I.R.E.

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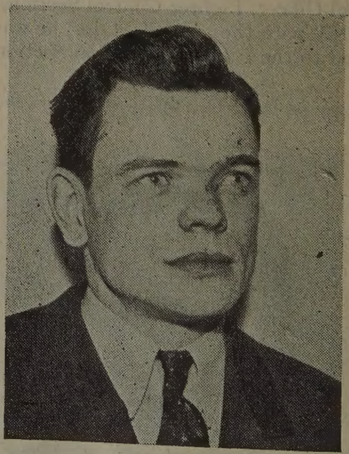
James N. Van Scoyoc (A'44) was born in Rockville, Indiana, on December 26, 1911. He received the B.S. degree in 1934 from Purdue University and did one year of graduate work in physics there in 1935. He was engaged in radio sales and service work until June 1942; and from then until the present time as instructor and laboratory supervisor in electronics at Illinois Institute of Technology. Since 1942 a major portion of his work has been the design and development of special laboratory and testing equipment. Mr. Van Scoyoc is a member of Sigma Pi Sigma.



James N. Van Scoyoc